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APPLICATION NUMBER: *60/350,414*
FILING DATE: *January 18, 2002*
RELATED PCT APPLICATION NUMBER: *PCT/US03/01854*

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J1057 U.S. PTO

PROVISIONAL APPLICATION COVER SHEET

This is a request for filing a PROVISIONAL APPLICATION under 37 CFR 1.53 (b)(2).

Docket Number	20191 PROV	Type a plus sign (+) inside this box→	+

INVENTOR(s)/APPLICANT(s)			
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J1002 U.S. PTO 60/350414 01/18/02

TITLE OF THE INVENTION (280 characters max)

ANALOG MODAMP

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ENCLOSED APPLICATION PARTS *(check all that apply)*

☒ Specification (Number of Pages) 52
☒ Small Entity Status is claimed
☒ Drawing(s) (included in spec.)
☐ Other (specify)

METHOD OF PAYMENT *(check one)*

☒ A check or money order is enclosed to cover the Provisional filing fees
☐ The commissioner is hereby authorized to charge filing fees and credit Deposit Account Number:

PROVISIONAL FILING FEE AMOUNT ($): 80.00

The invention was made by an agency of the United States Government or under a contract with an agency of the United States Government.

☒ No.
☐ Yes, the name of the U.S. Government agency and the Government contract number are: ____________

Respectfully submitted,

SIGNATURE [signature] Date Jan 18, 2002

TYPED or PRINTED NAME CLIFTON W. THOMPSON REGISTRATION NO. (if appropriate) 36,947

☐ Additional inventors are being named on separately numbered sheets attached hereto

PROVISIONAL APPLICATION FILING ONLY

2037D 11405 U

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Date of Deposit: January 18, 2002

I hereby certify that the enclosed Provisional Patent Application consisting of a cover sheet, and 52 pages of application, including drawings, and a return postcard, --is being deposited with the United States Postal Service "Express Mail Post Office to Addressee" on the date indicated above in an envelope addressed to the Assistant Commissioner for Patents, **Box New Provisional Patent Application**, General Delivery, Washington, D.C. 20231.

Respectfully submitted,

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Analog ModAmp
Theory of Operation

Michael E. Spencer, Ph.D.

ATC CONFIDENTIAL

November 5, 2001

Summary

This document describes the theory of operation and the circuit operation details for the Analog ModAmp realization of the ATC/SPS Project #1 as delivered last week.

The system block diagrams are presented for the generalized ModAmp and simulation results are shown for the Bi-Level SSB modulator/amplifier. The Bi-Level SSB modulator described below is an extension of the Bi-Level modulator described in the background document: "Description of Invention: Switched Mode Modulator and Amplifier (ModAmp)" dated June, 2001.

System Level Theory of Operation

Tri-Level and Bi-Level SSB

Figure 1 shows the signals for single sideband (SSB) synthesis. In this case, sine and cosine waves are compared to the input signals consisting of in-phase and quadrature components (90 degrees out of phase from one another). The in-phase and quadrature components are typically derived by using a Hilbert transformer. Fourteen events are defined in the figure at crossings of the input signals with the sine waves or at zero crossings of the sine waves.

Figure 2 shows the top-level block diagram of the modulator/amplifier. Sinusoidal input test signals are generated in quadrature and drive the real and imaginary inputs of the Event Generator block, detailed in Figure 3. The event generator derives a series of 14 short pulses, or event triggers, that are used by the AM and SSB modulators. The AM and SSB modulators are detailed in Figure 4 and Figure 5, respectively.

For binary output, we want the outputs to take on only two levels. The tri-level AM and SSB waveforms can be converted to bi-level waveforms by adding a square wave at the carrier frequency, as shown at the bottom of Figure 1. Alternatively, the bi-Level signal labeled as A'_S (the bold waveform) in Figure 1 yields the desired SSB output. The bi-Level output signal allows an amplifier to be built using a half bridge (instead of a full H-bridge) requiring only two output transistors.

A simulation was run with two test tones using the parameter settings shown in Figure 6. The top of Figure 7 shows the time domain waveforms for tri-level and A'_s-bi-Level modulation. The remainder of Figure 7 shows the spectrum for the two cases. Notice that the A'_s-bi-Level modulation has out-of-band signals centered around 80KHz. Since the PVDF transducers used in our application have negligible output above 55KHz, we use the simple A'_s-bi-Level modulation scheme in our Analog ModAmp.

The power supply noise/ripple rejection approach of Figure 8 (b) is used in the Analog ModAmp. It was shown in the invention disclosure document that if the reference oscillators' amplitudes are controlled in proportion to the MOSFET power supply voltage, then we achieve an output that is independent of the supply voltage.

Modulator Characteristics

Table 1 summarizes the characteristics of the various modulators. We assume an H-bridge requires 4 MOSFETs and a half-bridge requires 2 MOSFETs. The "transitions per carrier period" indicate the number of signal transitions of the modulator output per carrier period. The fewer transitions generally yield higher efficiency amplifiers.

Item 5 uses two bi-level AM modulators and takes the difference to synthesize the SSB output. Item 6 starts with "bi-level AM" and adds a square wave at the 3rd harmonic of the carrier. This is used to reduce the amplitude of the 3rd harmonic in the modulator output. Schemes that add higher order harmonics is also feasible. Item 7 starts with "tri-level AM" and adds a square wave at the 3rd harmonic of the carrier. Item 8 combines two "bi-level AM reduced 3rd harmonic" (item 6) to synthesize the SSB output. Item 9 is the new technique used in the Analog ModAmp described in this document.

Table 1: Characteristics of the various modulators

	Modulation Technique	# of Levels seen by load	# MOSFET switches required	transitions per carrier period	(transitions/period) per switch pair	
1	Tri-Level AM	3	4	4	2	
2	Tri-Level SSB	3	4	8	4	
3	Bi-Level AM	2	2	6	6	
4	Bi-Level SSB	2	2	10	10	
5	SSB (Bi-Level AM x 2)	3	4	12	6	Two independent Bi-Level AM modulators
6	Tri-Level AM (reduced 3rd harmonic)	3	4	8	4	This is Bi-Level AM with the 3rd harmonic square wave added.
7	Bi-Level AM (reduced 3rd harmonic)	2	2	10	10	This is Tri-Level AM with the 3rd harmonic square wave added.
8	SSB (Bi Level AM reduced 3rd x 2)	3	4	20	10	
9	Bi-Level SSB base on A' drive signal	2	2	4	4	This technique first described in this document. Has 80KHz noise component (for 40KHz carrier).

Pre-Processing Software for the Analog ModAmp

The source audio material is processed on a computer to generate an I (in-phase) and Q (quadrature) signals that are saved on the MP3 players right and left channels, respectively. The software is written in MATLAB/Simulink and the block diagrams are shown in Figure 9, Figure 10 and Figure 11.

Analog ModAmp Circuit Description

This amplifier accepts analog I and Q signals from an MP3 player. These signals are DC coupled so the carrier level can be controlled dynamically from the pre-processor software output. An MP3 player must be used that preserves the DC term. The Samsung YP-30S was selected for this application.

The Analog ModAmp was designed and built that performs bi-level lower-sideband modulation. The circuits are described in the following subsections.

Power Supplies

The power supplies are shown in the schematic in Figure 12. The high voltage power supply is a simple off-line supply consisting of EMI filter components, a full wave bridge rectifier and filter capacitors. This high voltage supply powers the main MOSFETs and the auxiliary power supply. The auxiliary supply produces 5V for the ModAmp's control circuitry and the MP3 player. It is based on the Power Integrations TNY264 fly-back regulator chip. This chip packs all the control logic and the main switch for the complete transformer isolated fly-back power supply.

Sine/Cosine Reference Oscillators

The master clock oscillator is the LTC1799 chip, U5, in Figure 13. The inductor L7 ensures that oscillator noise is not conducted into the 5V supply. The oscillator runs at four times the carrier frequency. A simple state machine consisting of two D-flip-flops, U6, generates a pair of square waves in quadrature at the carrier frequency. Two 4th order Chebyshev lowpass filters consisting of U7, U8 and associated R's and C's, are used to filter the harmonics of the square waves, leaving nearly pure sinusoidal tones. The result is the sine and cosine carrier reference signals. DC blocking capacitors C17 and C25 decoupled the accumulated offset error in the filter. A final gain stage, U9, is used to boost the signal level.

Power Supply Rejection Circuit

The circuits in the lower half of Figure 13 consisting of U10 and U11 are the power supply rejection circuits. These circuits force a symmetric power supply voltage across the D-Flip-flops that is proportional to the high voltage power supply. Since the outputs of the D-Flip-flops swing from rail-to-rail, the quadrature square wave output amplitudes are proportional to the high voltage supply. And finally, the sine and cosine output amplitudes will be proportional to the high voltage supply, as desired.

The circuit gains are set such that the output sine waves will clip with high input line voltages. This clipping has no negative consequence since the peaks of the waveforms are not used by the subsequent comparator circuits. The increased amplitude in the design increases the overall dynamic range.

Reset/Brown-Out Protection Circuit

The reset chip U22 triggers a 2 second active low reset pulse on power up. If the VCC input drops below 4.00V reset is also asserted. The RESET_F signal, when active, disables the high voltage power supply to the main MOSFETs through the switch, Qa in the power supply schematic, Figure 12. This system is used to delay the power-up of the main MOSFETs until the 5V supply has stabilized and the MOSFET drive signals are valid.

This circuit also shuts down the MOSFET power supply under brown-out conditions where the AC line is below about 80VAC. The design performs the brown-out reset using the diodes D4 to drag down the reset chip's VCC input when the high voltage supply droops.

This reset behavior is critical since the sine wave reference signals are proportional to the high voltage and lowered reference signals would eventually lead to invalid control waveforms at the comparator outputs.

MP3 Player Interface and Dynamic Carrier Circuits

The inputs to the ModAmp consists of in-phase and quadrature audio signals with DC controlling the carrier level. The circuits of U12, U17 and U14B in Figure 14 take the MP3 player input and generate a hard-limited (0 to 5V) signal the is nominally at +5V/2 (2.5V) with no input signal. Clipping excessive input levels is critical at this stage to avoid invalid comparator outputs in the subsequent circuits.

The op-amp circuit design takes the ratiometric signal from the MP3 player output (the DAC in the player is proportional to the 2.47V reference times the digital code) and generates an output that drops below the nominal 2.5V as the carrier level increases. The circuit is designed to cancel the offset voltage errors that would normally occur with +5V power supply variations.

LPF and Comparator Circuits

The 15KHz 2nd order Bessel lowpass filter U14A and U19A removes high frequency signals that may be in the input and drives the negative comparator inputs. The top comparator output has falling and rising edges that correspond to the times that the output MOSFETs should be cleared. See the wave forms in Figure 1 (I and Q are flipped so the ModAmp performs lower sideband modulation). Similarly the lower comparator's edges correspond to times when the output MOSFETs should be set.

Pulse Synthesizer Circuits

The outputs of the comparators are fed to an edge detector circuit that generates a short 350nS pulse on both the positive going and negative going comparator transitions. These "event trigger" signals are used to set and clear the output state.

Pulse Driver Circuits

Figure 15 shows the A_set and A_clear event trigger pulses controlling the gates of Q1 and Q2. Next, Q1 and Q2 drives T3 in push-pull mode. The secondaries of T3 have a series of short alternating negative and positive going pulses with an amplitude of about +/- 10V.

MOSFET Pulse to Level Converters and Output Stage

For each main MOSFET is driven by a pair of steering MOSFETs (Q3,Q4 and Q5,Q6) which converts the short event trigger pulses to steady-state voltage levels. By using MOSFETs with different gate thresholds and the secondary-to-secondary coupling in the pulse transformer, this novel circuit design guarantees that the main MOSFETs will avoid cross-conduction (or shoot through) and will operate over a wide duty cycle range.

Transducer Isolation and Matching Stage

Transformer T4 achieves the required transducer isolation from the mains and the matching inductor and capacitor form a tuned circuit with the transducer to help boost the voltage level and equalize the system. This area of the circuit is sub-optimal and still needs some work.

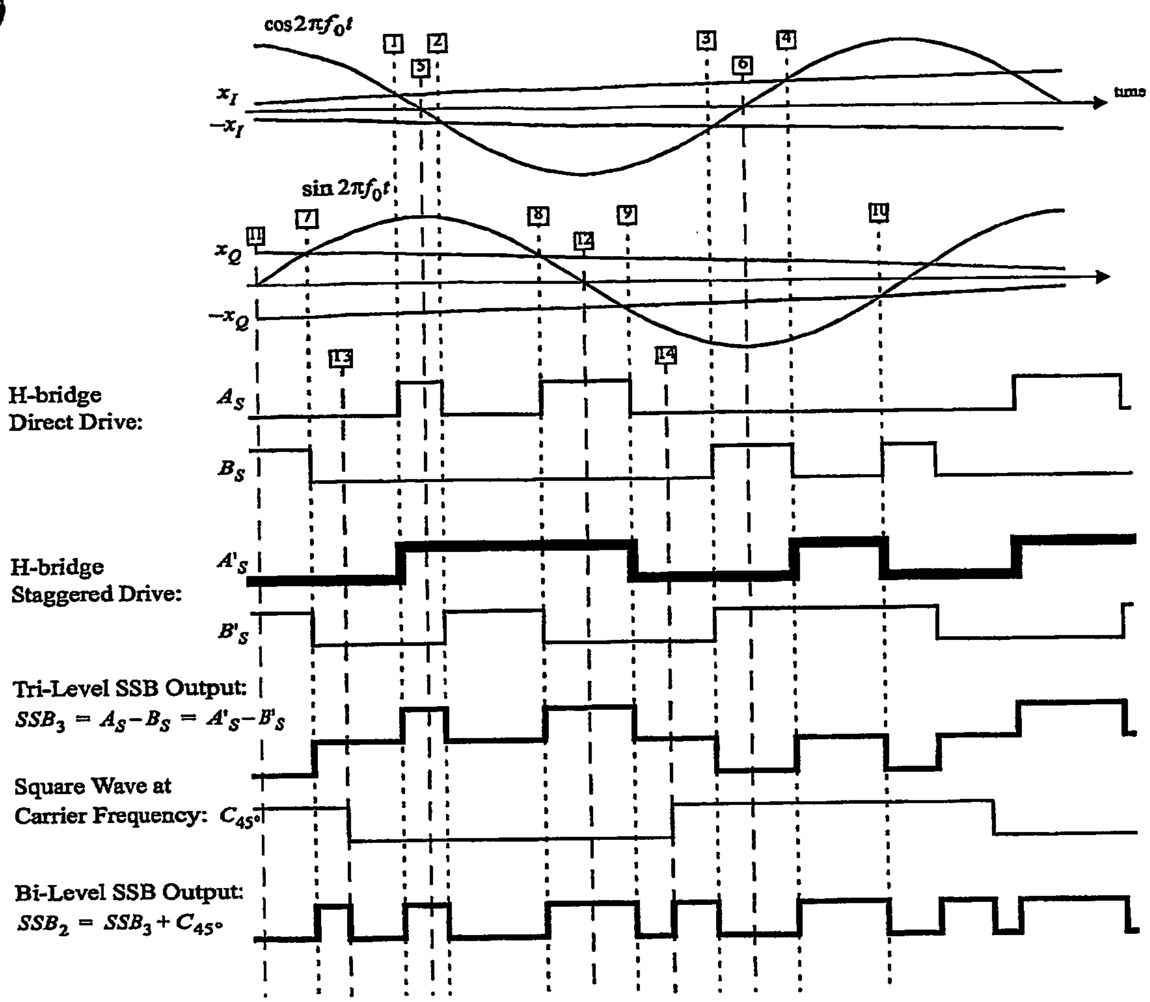


Figure 1: Sine/Cosine wave synthesis of the Tri-Level and Bi-Level SSB signals. The in-phase component of the input signal is x_I, and the 90-degree shifted quadrature component is x_Q. Two alternative sets of H-bridge drive signals are shown: A_S, B_S and A'_S, B'_S. If the load is placed in the center of the H-bridge, the differential gives us the desired Tri-Level SSB Output signal. The numbered boxes label the timing events that trigger the waveform transitions. The Bi-Level Output is derived by adding a square wave at the carrier frequency. The **Bolded** signal is delivered to the load in the Analog ModAmp.

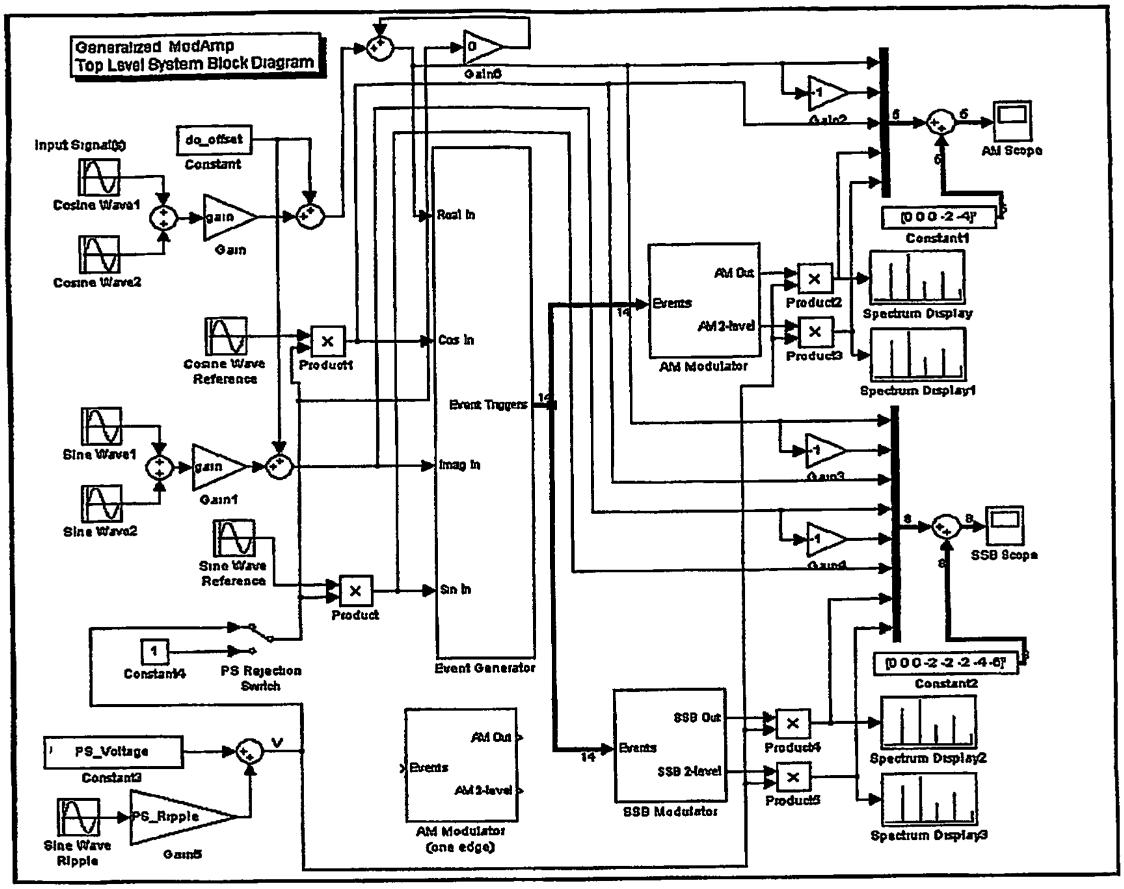


Figure 2: ModAmp top-level block diagram shows AM and SSB modulators based on Sine/Cosine wave synthesis. The Event Generator, AM Modulator, SSB Modulator blocks are detailed in the next three figures.

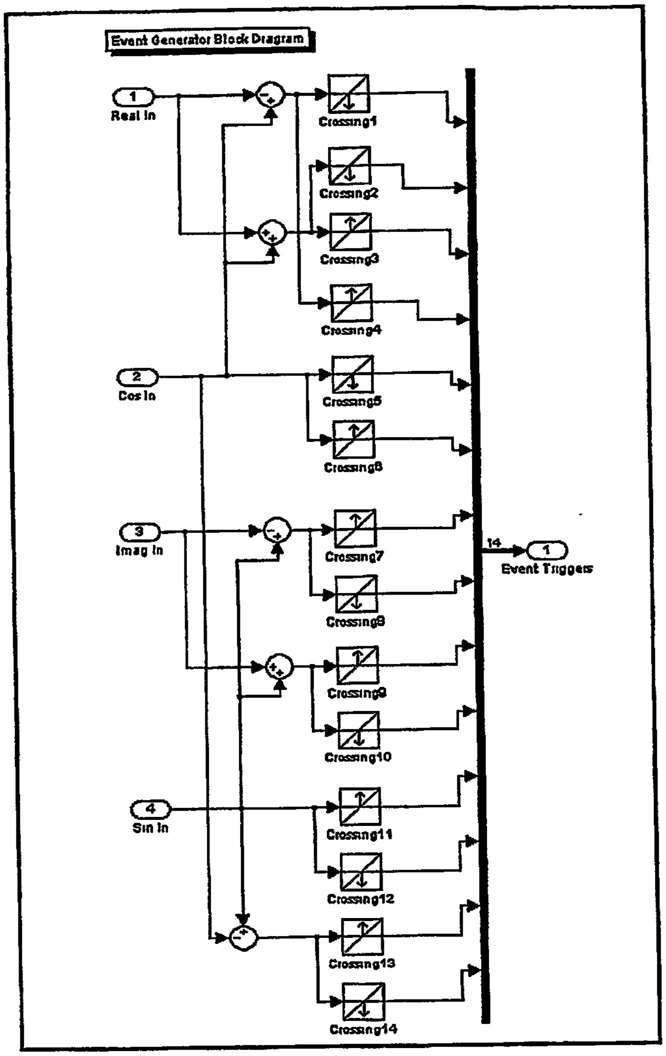


Figure 3: Event generator block diagram for ModAmp. Each zero crossing detector outputs a short pulse when the input crosses zero in the direction shown. These event trigger signals, from top to bottom, corresponding to the event numbers in Figure 1.

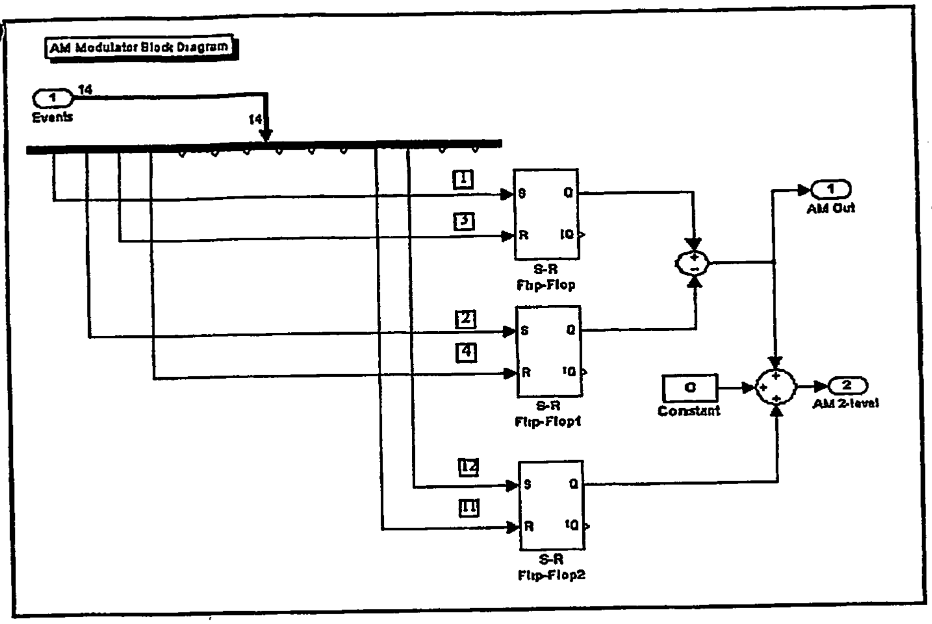


Figure 4: AM Modulator Block for Generalized ModAmp. The event triggers set and reset the filp-flops to generate the Tri-Level and Bi-Level AM outputs using staggered drive.

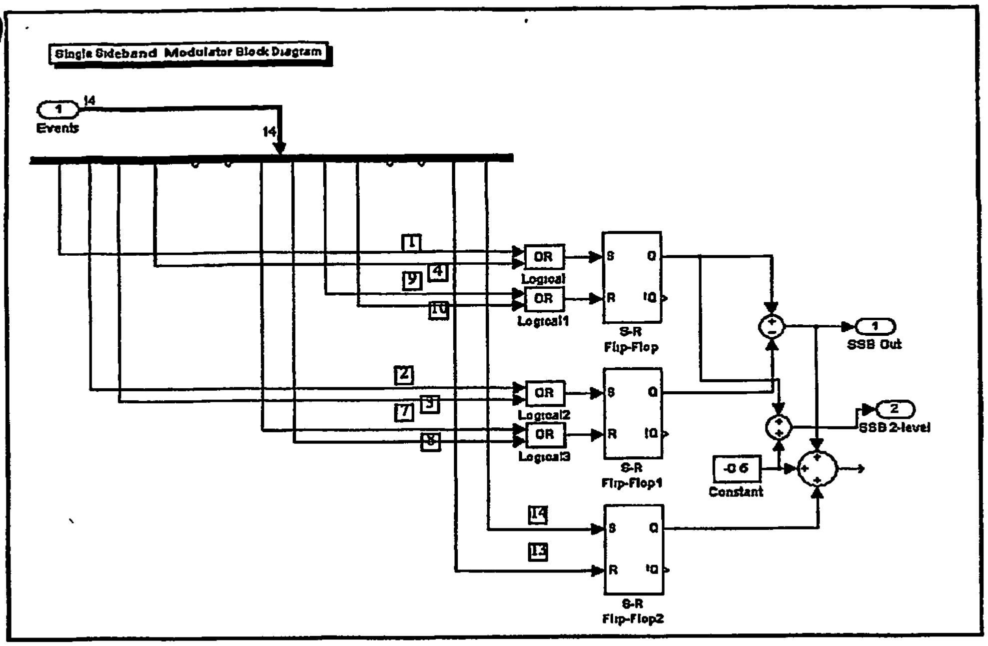


Figure 5: SSB Modulator Block for Generalized ModAmp. The event triggers set and reset the filp-flops to generate the Tri-Level and Bi-Level SSB outputs using staggered drive as in Figure 1. The SSB 2-Level output is simply the A' drive signal.

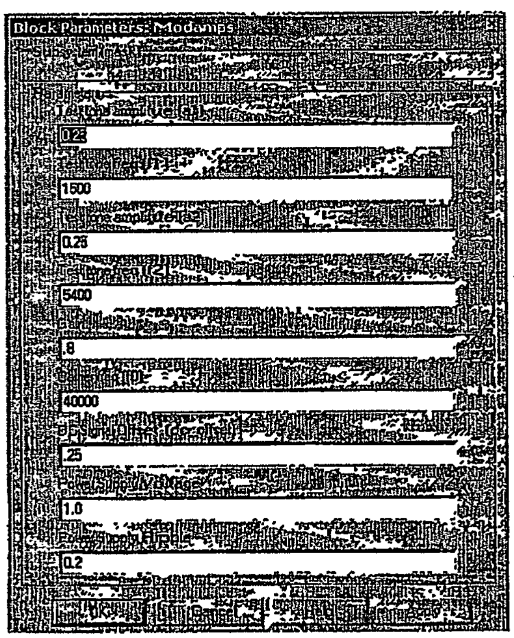


Figure 6: Settings for simulation shown in Figure 7.

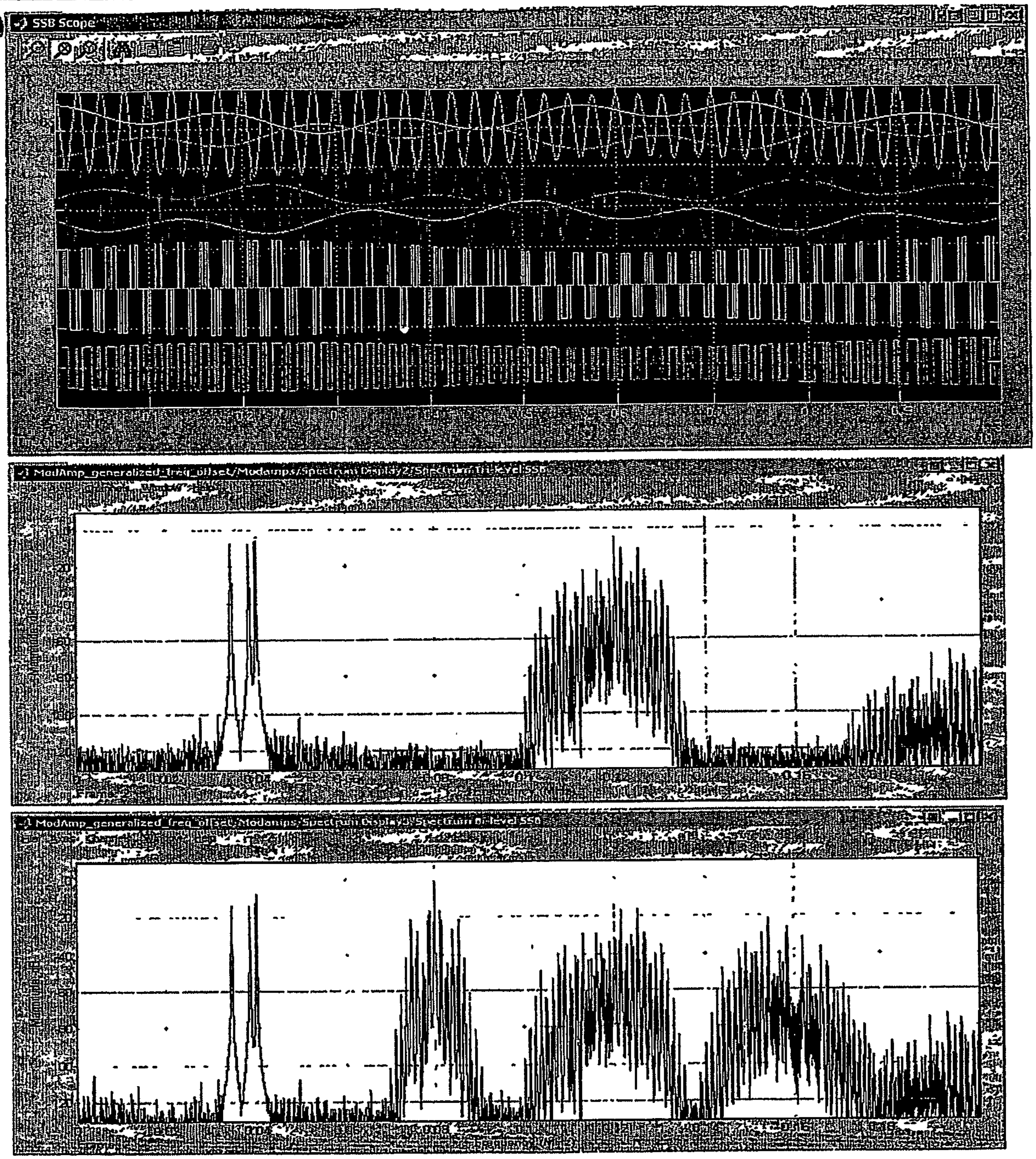


Figure 7: This simulation compares SSB modulators with Tri-Level (top spectrum) and Bi-Level (based on the Staggered Drive signal: A', bottom spectrum). The analog ModAmp implements the Bi-Level case. The results were generated by the ModAmp system depicted in Figure 2.

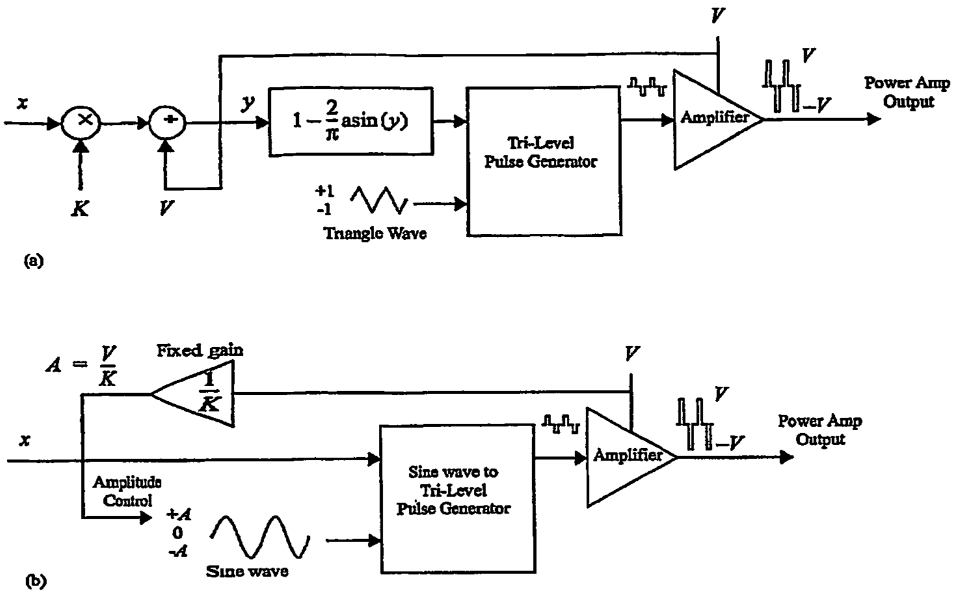


Figure 8: Power supply noise rejection may be achieved by the feedforward technique as in (a) for the triangle wave based tri-level pulse generator, or as in (b) for the sine wave based modulator. As the power supply voltage, V, changes the pulse-widths are appropriately adjusted to maintain a consistent output.

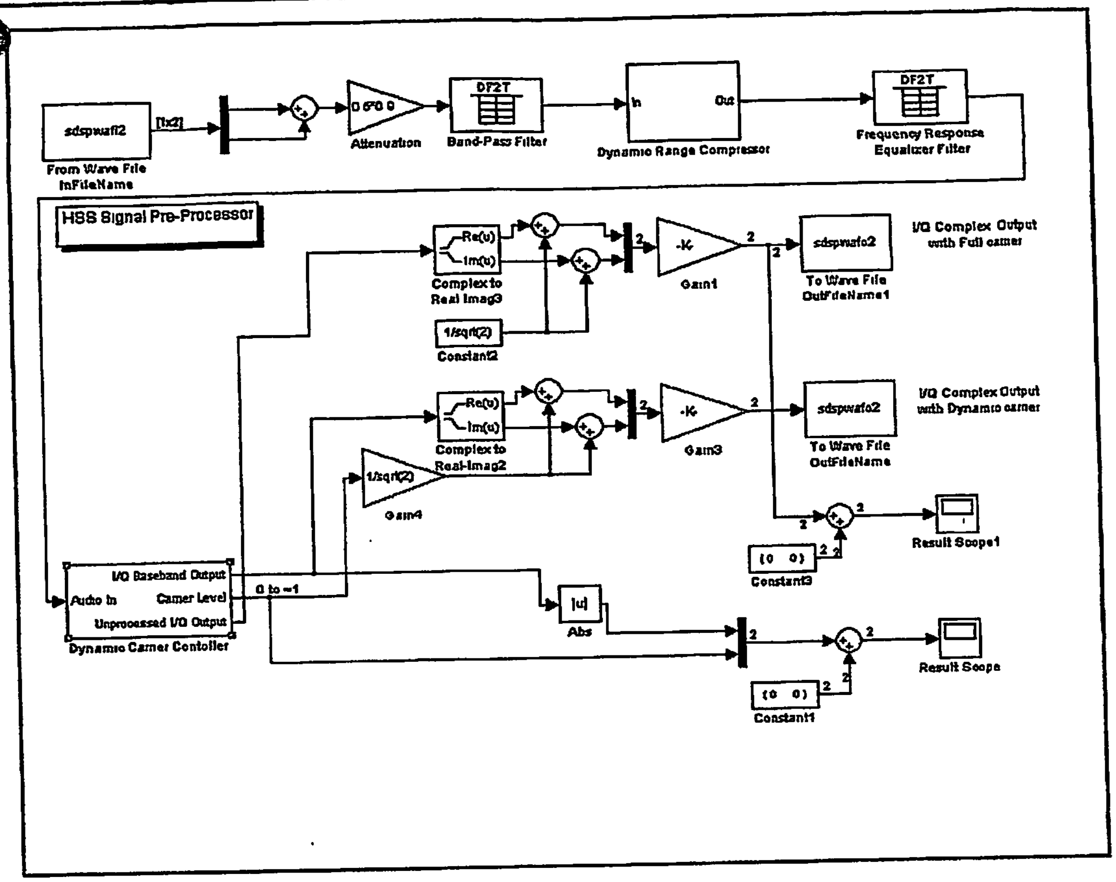


Figure 9: Pre-processor software top-level block diagram shows that an input wave file is processed into an output wave file. The Dynamic Range Compressor and Dynamic Carrier Controller blocks are detailed in the next two figures.

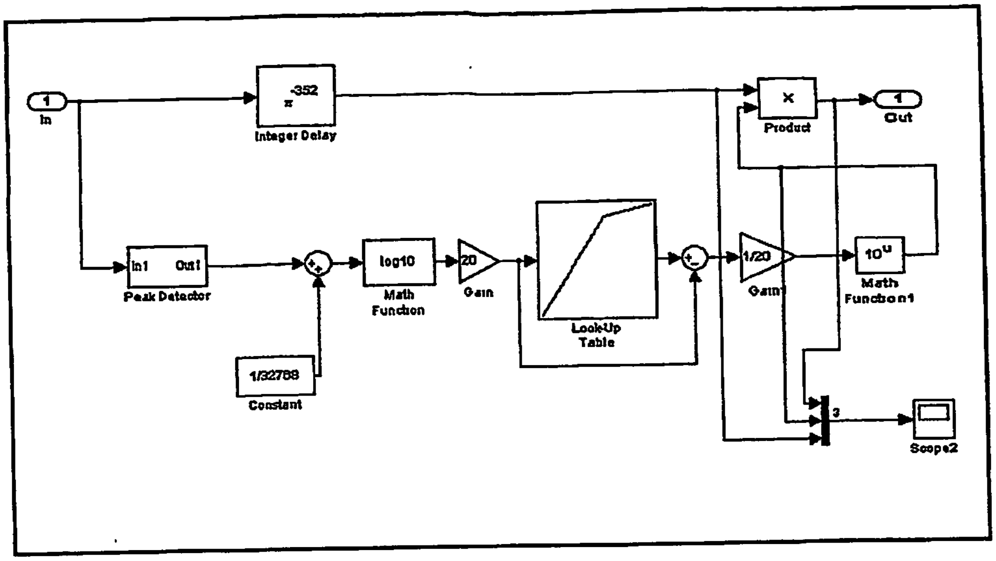


Figure 10: Dynamic Range Compressor

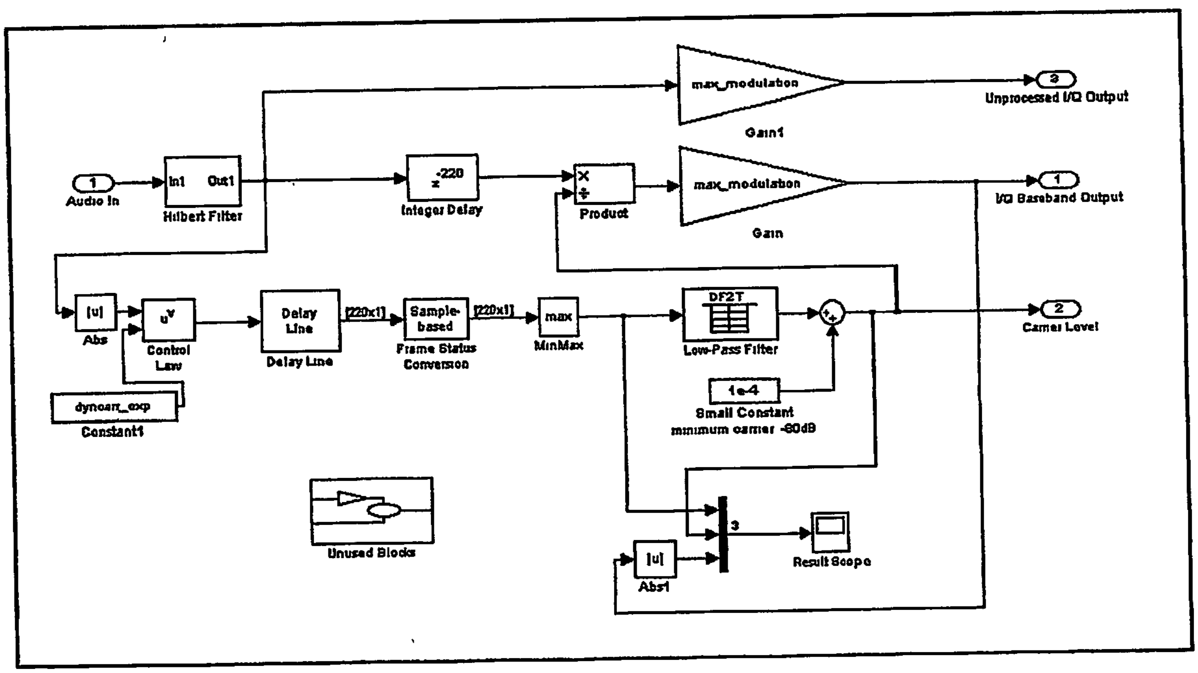


Figure 11: Dynamic Carrier Controller.

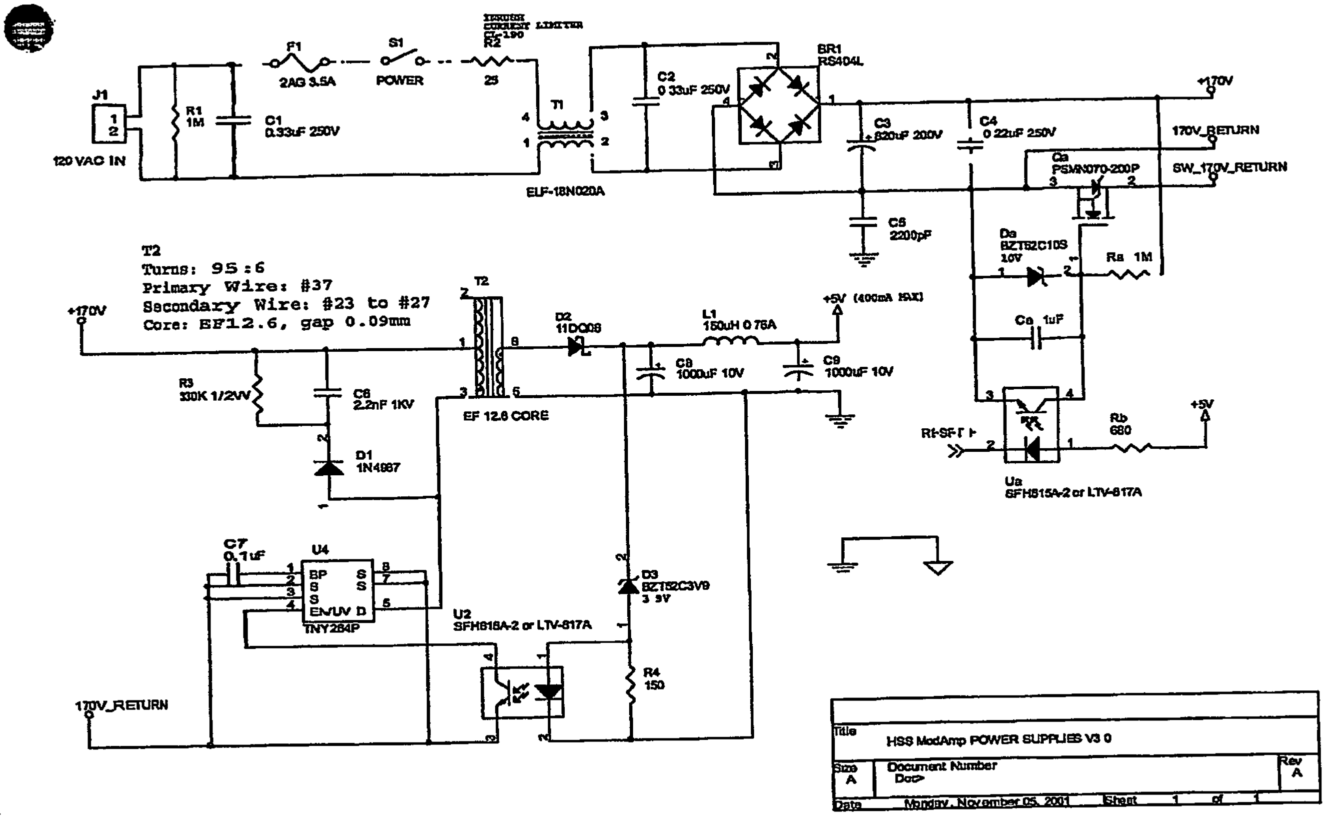


Figure 12: Analog ModAmp, page 1.

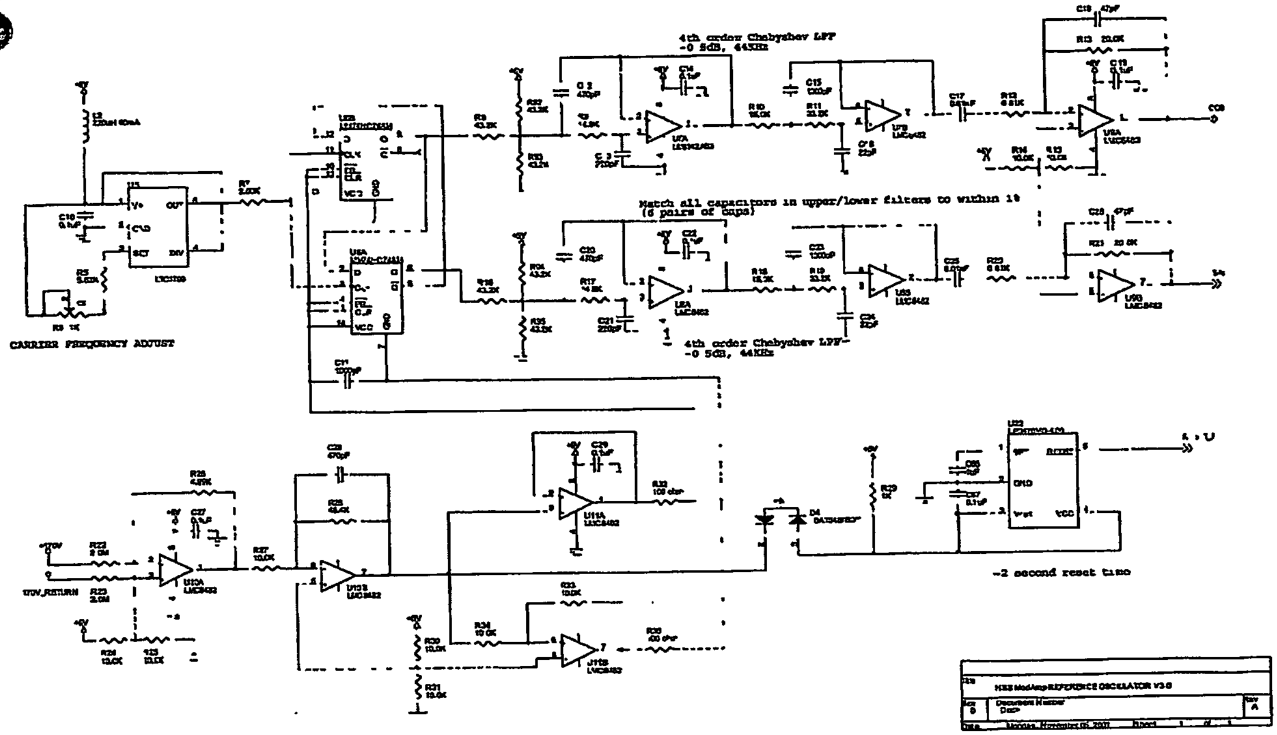


Figure 13: Analog ModAmp, page 2.

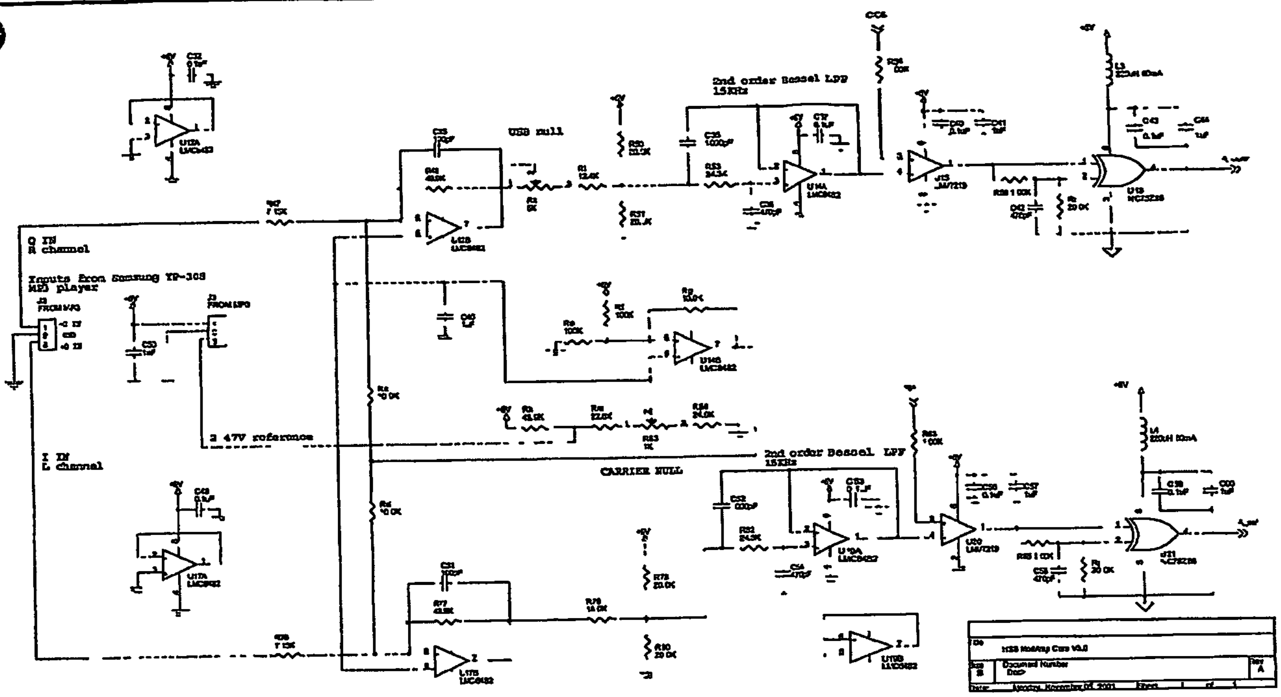
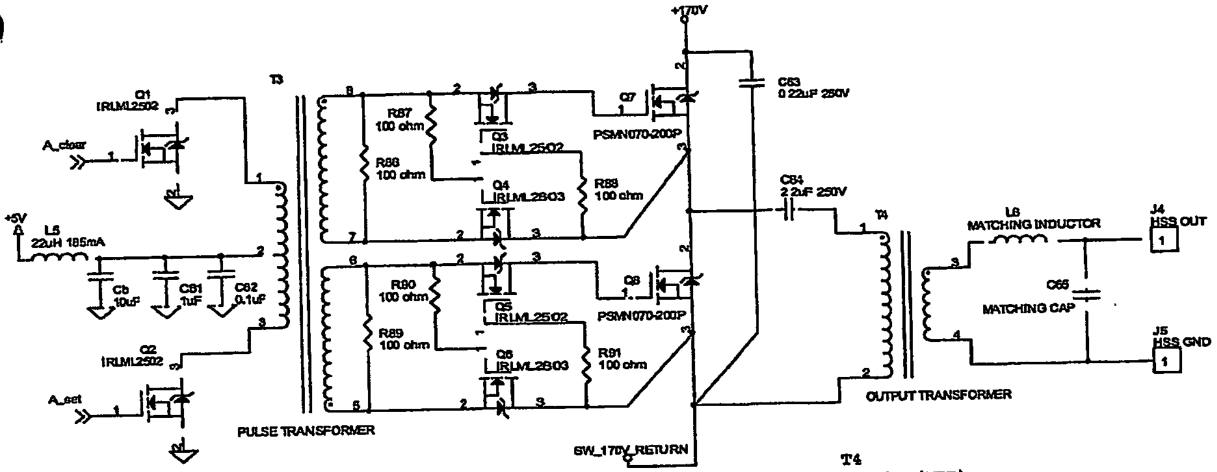


Figure 14: Analog ModAmp, page 3.



T4
50:~24 (TBD)
Primary Wire: #18 enamel
Secondary Wire: #18 enamel
Core: ZP Toroids 42915-TC (gray core)

T3
Turns: 5-5: 11:11
Primary Wire: #36 double-insulated (Rubadue D32A01T09-1.5))
Secondary Wire: #34 enamel
Core: Lodestone Pacific YW-40502-TC
Core Carrier: Lodestone Pacific STM180-08/100

Title	HSS ModAmp GATE DRIVER & OUTPUT STAGE V3.0	
Size A	Document Number Doc>	Rev A
Date	Monday, November 05, 2001	Sheet 1 of 1

Figure 15: Analog ModAmp, page 4.

	A	B	C	D	E	F	G	H	I	J
3	**Power Supply Input**									
4	VACMIN	Volts	85							Minimum AC Input Voltage
5	VACMAX	Volts	265							Maximum AC Input Voltage
6	FL	Hertz	50							AC Main Frequency
7	TC	mSeconds	2.30							Bridge Rectifier Conduction Time Estimate
8	Z		0.60							Loss Allocation Factor
9	N	%	72.0							Efficiency Estimate
10										
11	**Power Supply Outputs**									
12	VOx	Volts		5.00						Output Voltage
13	IOx	Amps		0.400						Power Supply Output Current
14										
15	**Device Variables**									
16	Device		TNY264							Device Name
17	PO	Watts	2.00							Total Output Power. Comment: Overload Po has been increased to compensate for device current limit delay (tILD). Po = 3.82 W
18	VDRAIN	Volts	577							Maximum Drain Voltage Estimate (Includes Effect of Leakage Inductance)
19	VDS	Volts	6.7							Device On-State Drain to Source Voltage
20	FSnom	Hertz	132000							TinySwitch-II Switching Frequency
21	FSmin	Hertz	120000							TinySwitch-II Minimum Switching Frequency (Inc. Jitter). Comment: Full load operating frequency is less than minimum. Fs = 63KHz
22	FSmax	Hertz	144000							TinySwitch-II Maximum Switching Frequency (Inc. Jitter)
23	KRPKDP		1.14							Ripple to Peak Current Ratio
24	ILIMITMIN	Amps	0.23							Device Current Limit, Minimum
25	ILIMITMAX	Amps	0.27							Device Current Limit, Maximum. Comment: Actual current limit may exceed ILIMITMAX and is limited to a safe level. Due to effects of device current limit delay (tILD). ILIMITMAX = 0.32 A
26	IRMS	Amps	0.10							Primary RMS Current
27	DMAX		0.48							Maximum Duty Cycle
28										
29	**Power Supply Components Selection**									
30	CIN	uFarads	6.8							Input Filter Capacitor
31	VMIN	Volts	90							Minimum DC Input Voltage
32	VMAX	Volts	375							Maximum DC Input Voltage
33	VCLO	Volts	130							Clamp Zener Voltage
34	PZ	W	0.2							Estimated Primary Zener Clamp Loss
35										
36	**Power Supply Output Parameters**									
37	VDx	Volts		0.5						Output Winding Diode Forward Voltage Drop
38	PIVSx	Volts		29						Output Rectifier Maximum Peak Inverse Voltage
39	ISPx	Amps		2.71						Peak Secondary Current
40	ISRMSx	Amps		1.06						Secondary RMS Current
41	IRIPPLEx	Amps		0.98						Output Capacitor RMS Ripple Current

	A	B	C	D	E	F	G	H	I	J
43	**Transformer Construction Parameters**									
44	Core/Bobbin		EF12.6							Core and Bobbin Type
45	Core Manuf.		Generic							Core Manufacturing
46	Bobbin Manuf.		Generic							Bobbin Manufacturing
47	LPmin	uHenries	1297							Minimum Primary Inductance
48	NP		95							Primary Winding Number of Turns
49	AWG	AWG	37							**Primary Wire Gauge (Rounded to next smaller standard AWG value). Warning! Wire gauge (AWG) greater than recommended maximum (36 AWG),and may not be manufacturable.**
50	CMA	Cmils/A	205							Primary Winding Current Capacity (200 CMA 500)
51	VOR	Volts	86.94							Reflected Output Voltage
52	BW	mm	7.50							Bobbin Physical Winding Width
53	M	mm	0.0							Safety Margin Width
54	L		2.0							Number of Primary Layers
55	AE	cm^2	0.13							Core Effective Cross Section Area
56	ALG	nH/T^2	144							Gapped Core Effective Inductance
57	BM	Gauss	2944							Maximum Operating Flux Density
58	BAC	Gauss	1380							AC Flux Density
59	LG	mm	**0.09**							**Gap Length (Lg > 0.051 for TOP22X, Lg > 0.1 for TOP23X). Warning! Gap length (LG) is smaller than the practical limit. Increase transformer size, increase NS, increase KrpKdp.**
60	LL	uH	25.9							Estimated Transformer Primary Leakage Inductance
61	LSEC	nH	20							Estimated Secondary Trace Inductance
62										
63	**Secondary Parameters**									
64	NSx			6.00						Secondary Number of Turns
65	Rounded Down NSx									Rounded to Integer Secondary Number of Turns
66	Rounded Down Vox	Volts								Auxiliary Output Voltage for Rounded to Integer NSx
67	Rounded Up NSx									Rounded to Next Integer Secondary Number of Turns
68	Rounded Up Vox	Volts								Auxiliary Output Voltage for Rounded to Next Integer NSx
69	AWGSx Range	AWG		23 - 27						Secondary Wire Gauge Range (CMA range 500 - 200). Wire gauge (AWG) is less than 26 AWG. Consider parallel winding (see AN-18, AN-22).

Table 2: This Table is the specification of the T2 and was generated by Power Integrations Pi Expert Software for the 5V, 400mA power supply.

Description of Invention:
Switched Mode Modulator and Amplifier (ModAmp)

Michael E. Spencer, Ph.D.

June, 2001

Abstract

CONFIDENTIAL

Novel methods and apparatus are described for the modulation and amplification of signals. An input signal can be used to modulate a carrier with a variety of schemes such as amplitude modulation (AM) or single sideband modulation (SSB). The modulator generates an output with a small number of discrete output amplitudes (or voltage levels). Typically 2 or 3 discrete output levels are used. The modulator output can be amplified to any arbitrary level by increasing the voltage swing. In its simplest form, the modulator output is a binary signal that is either low or high. This binary signal can be applied to MOSFET switches to increase the voltage swing thereby increasing or amplifying the signal. By using this switching technique, high-efficiency power modulator/amplifiers may be realized. The combined modulator/amplifier is referred to as a ModAmp in this white paper. The ModAmp output spectrum consists of the desired modulated signal plus high frequency switching products. In typical applications, a lowpass filter is used to remove the high frequency switching products. It is not necessary to have a carrier tone present in the modulator output. The AM or SSB signal may have a carrier present or may be operated in a suppressed carrier fashion. When the carrier is suppressed, the SSB ModAmp performs frequency translation and amplification. That is, the input signal is frequency shifted and amplified by the ModAmp.

Index Terms -

ModAmp; modulator; amplifier; modulator/amplifier; amplitude modulation; AM modulation; single-sideband modulation; SSB modulation; upper sideband modulation; lower sideband modulation; suppressed carrier; band-pass amplifier; binary AM; binary SSB; frequency shifter; frequency translation; bi-level modulation; tri-level modulation; power modulator; power amplifiers; high efficiency; low distortion; pulse width modulation; PWM; pulse duration modulation; PDM; switching amplifier; switch mode; Class-D amplifier; PWM power amplifier; feedforward; feed-forward; power supply rejection; PCM to PWM converter; internal modulator; DSP algorithms; harmonic distortion; signal processing; parametric arrays; AM transmitters; SSB transmitters; medical ultrasound; SONAR

Applications

Introduction to Parametric Arrays-- One application of the ModAmp is the parametric array. A parametric array system generates audible sound in air using a transducer that emits only inaudible ultrasonic energy. The non-linear acoustic properties of the air perform demodulation of the ultrasonic signal to generate the audible sound.

In a typical parametric array system, an ultrasonic carrier signal is modulated by audio source, then it is amplified and applied to a high-intensity speaker, emitter or transducer. The intense ultrasound causes the air in front of the emitter to exhibit a non-linear transfer characteristic. The air's nonlinearity generates intermodulation distortion products in the form of sum and difference frequencies. The audible sound is produced by the difference frequencies. See Figure 1. For example, an upper sideband modulator with a 1 kHz input and a 40 kHz carrier frequency will produce a 40 kHz and a 41 kHz signal at the speaker. The non-linear air column demodulation will produce an audible 1 kHz tone.

Parametric arrays give the ability to direct or focus sound into tight beams using physically small transducers. The parametric array concept works both in gases such as air or in fluids such as water. The ModAmp is an ideal solution for the modulation and amplification functions required in parametric array systems because of its small size and high-efficiency.

Other applications include AM and SSB transmitters, SONAR signal modulation and amplification, medical ultrasound applications, frequency translating amplifiers, bandpass amplifiers, etc.

Waveform Synthesis

By varying the positive and negative pulse widths of the tri-level signal shown in Figure 2, we can change the intensity of the fundamental tone, a_1. Pulse width modulation of this tri-level waveform is an effective way to perform AM modulation at the carrier frequency. But first, we must understand how the fundamental tone amplitude varies with pulse width. The amplitude of the fundamental and harmonic tones are derived in the next section. It is shown that the tri-level waveforms generate the fundamental and only odd harmonics. This puts the first out-of-band component at three times the carrier frequency.

We have described a tri-level waveform, however, a bi-level (binary) waveform may also be used. This is detailed in a later section. These modulation techniques are easily extended to multi-level signals (e.g. 4-level, 5-level, etc.).

Amplitude of Fundamental

It is shown in this section, that the fundamental amplitude of the output signal is a non-linear function of the pulse width parameter. Recall that a real periodic signal with period $T = 1/f_0$ can be written in terms of it's Fourier Series expansion:

$$f(t) = \frac{a_0}{2} + \sum_{n=1}^{\infty} (a_n \cos n\omega_0 t + b_n \sin n\omega_0 t) \tag{1}$$

where

$$a_n = \frac{2}{T}\int_{t_1}^{(t_1+T)} f(t) \cos n\omega_0 t dt \tag{2}$$

and

$$b_n = \frac{2}{T}\int_{t_1}^{(t_1+T)} f(t) \sin n\omega_0 t dt \tag{3}$$

with

$$\omega_0 = 2\pi f_0 = \frac{2\pi}{T}. \tag{4}$$

The Fourier coefficients, a_n, and b_n, represent the amplitudes of the cosine and sine signals, respectively, that make up the spectrum of the periodic signal $f(t)$. For the Bidirectional pulse waveforms in Figure 2, and using (2) and (3), it can be shown that the amplitudes of the fundamental and harmonic tones are given by

$$a_n = \begin{cases} V\frac{4}{\pi n}\sin(2\pi n f_0 \tau), & \text{for } n = 1, 3, 5, \dots \text{ odd integers} \\ 0, & \text{otherwise} \end{cases} \tag{5}$$

and

$$b_n = 0, \quad \text{for all } n \tag{6}$$

where V is the peak value of the output waveform (or the power supply voltage) as labeled in Figure 2.

Of particular interest is how the amplitude of the fundamental, a_1, varies with the pulse-width control parameter τ. From (5), it can be seen that the fundamental tone amplitude is given by the following non-linear function of τ:

$$a_1 = V\frac{4}{\pi}\sin\left(\frac{2\pi\tau}{T}\right). \tag{7}$$

The fundamental amplitude has a maximum peak value of $V(4/\pi) \cong V \times 1.273$ when $\tau = T/4$. This corresponds to a square wave output signal. Note that the peak output level at the carrier frequency can be greater than the power supply rail, V.

We can use a triangle wave and two comparators to generate the tri-level waveform as shown in Figure 3. From the figure, notice that as the comparator threshold variable y varies from 0 to 1, τ will vary from $T/4$ to 0, or

$$\tau = \frac{T}{4}(1-y). \tag{8}$$

Suppose we want to perform linear AM modulation, that is we want the output's fundamental tone amplitude to vary **linearly** (between 0 and it's maximum of $4/\pi$) with some input variable or signal, x. Specifically, let the amplitude of the fundamental be

$$a_1 = K\frac{4}{\pi}x \tag{9}$$

where K is a constant (typically $K = V$). If $K = V$ then the fundamental amplitude will vary between 0 and it's maximum of $V(4/\pi)$ as x varies from 0 to 1.

By combining (7), (8), and (9) it is easy to show that the comparator threshold variable y in Figure 3 can be written as

$$\boxed{y = 1 - \frac{2}{\pi}\operatorname{asin}\left(\left(\frac{K}{V}\right)x\right)}. \tag{10}$$

Equation (10) is a necessary condition to achieve a linear variation of fundamental amplitude with the control variable, x. This equation applies when using the linear pulse width modulator of Figure 3. Equation (10) is a necessary (but not sufficient) condition for low distortion AM and SSB modulation.

Notes:

1. The spectrum given by the Fourier series expansion is only valid for a periodic signal, hence, it has been assumed the pulse width, (and τ) has remained constant with time.

2. Feeding the audio signal into a comparator with a triangle wave (a naturally sampled bi-pulse width generator) will cause significant distortion of the output spectra due to the non-linear relationship of pulse-width to

fundamental tone amplitude. Therefore, we must linearize the system with the arcsine function. Distortion will result if we don't perform the arcsine function with this modulator.

AM and SSB Modulation

Next, rather than holding the control variable, x, at a constant, we let it change with an input signal. This achieves AM modulation of the input signal at the carrier frequency of f_0. Figure 4 shows a block diagram of the AM and SSB modulators that use an arcsine linearizer and triangle wave comparator. The input consists of a sum of two sinusoides, one at 2,100Hz and the other at 9,300Hz. Next, a DC offset is added to the signal and it is passed into the AM linearizer. A a 40 kHz triangle wave is used with the tri-level pulse generator to generate the AM output.

To achieve SSB modulation, a similar signal processing path is used in the lower part of the block diagram, with the difference being the input signal is presented in quadrature and the triangle waveform is shifted by 90 degrees. For a generalized input signal a Hilbert transform can be used to generate the required analytic signal (consisting of in-phase and quadrature components). A lower sideband output is derived by subtracting the 2 AM outputs. Upper sideband would result if we added the two AM outputs.

The waveforms for the AM and SSB modulators are shown in Figure 5. In general, the SSB waveform can take on 5 different amplitude levels since it is the sum of two tri-level waveforms. However, if the modulator's offset constants are set appropriately and the input signal levels are limited to a certain value, then the SSB output will only take on only three levels.

The spectrum of the AM and SSB outputs are shown in Figure 6. In the AM case the carrier and 40 kHz and the upper and lower side tones are clearly seen on the left portion of the spectrum. Switching products centered at three times the carrier frequency (120 KHz) are also visible. The switching products may be filter out by a lowpass filter.

The spectrum displays are generated in the simulation by passing the analog signal through 150 kHz, 8th-order Elliptic lowpass filter, sampling the analog signal at 400 kHz and taking a 8,192 point Fast Fourier Transform (FFT) with a Hanning window. Higher frequency switching products are also present but are not shown above 200 KHz. The spectrum display is only accurate out to 150 KHz because this is where the lowpass filter begins to roll-off. The filter prevents aliasing in the FFT.

To realize the power amplifier portion of the ModAmp, the switching output waveforms are increased to the desired amplitude and passed through a lowpass filter to attenuate the high frequency switching components. A power amplifier based tri-level SSB modulation has been built using an H-bridge and appropriately switching the

two halves of the bridge to achieve the tri-level outputs. A ModAmp prototype was built that uses this technique and is detailed a letter section.

Note that the DC bias sets the nominal carrier level. In an alternative realization shown in a section below we can operate with the carrier suppressed.

Sine Wave Comparator

Above we computed the arcsine of the input signal, x so we could achieve linear operation when using a comparator with a triangle wave input (the linear bi-pulse width generator in Figure 3). Instead of a triangle wave, we can implicitly compute the arcsine by using a sine wave signal with a different comparator configuration. This eliminates the need for the arcsine function in the signal path.

A linearized tri-level output signal can be directly synthesized by using a sine wave reference oscillator as shown in Figure 7. In this case, a pulse is generated during the time that the absolute value of the sine wave is less than the input, x. The point at which the sine wave intersects the input value, x defines the time instant $t = \tau$, and the following equation can be deduced from the figure,

$$x = A\sin\left(\frac{2\pi}{T}\tau\right) \tag{11}$$

where A is the amplitude of the sine wave reference signal.

Solving this equation for τ and substituting it into the equation for the fundamental tone amplitude, (7), gives us a linear relationship between the input, x and the output tone amplitude:

$$a_1 = \left(\frac{V}{A}\right)\frac{4}{\pi}x. \tag{12}$$

If A is set to V/K then this equation is identical to (9). The important point here is that the fundamental amplitude is a linear function of the input or control variable, x.

Tri-Level AM and SSB with H-bridge

We can implement tri-level AM by using an H-bridge and appropriately driving the two half-bridges. Figure 8 shows the various waveforms that may be used to generate the AM modulator output. The two halves of the H-bridge can be driven by the A and B "direct drive" signals to achieve the tri-level AM output. Alternatively "staggered drive" may be used as shown in the A' and B' signals. All the output waveforms can be generated by

triggering state transitions on the timing events as labeled in Figure 8. For tri-level output, the DC bias must be appropriately set and input signal amplitude must be limited.

Figure 9 shows the signals for single sideband (SSB) synthesis. In this case, sine and cosine waves are compared to the input signals consisting of in-phase and quadrature components (90 degrees out of phase from one another). The in-phase and quadrature components are typically derived by using a Hilbert transformer. Fourteen events are defined in the figure at zero crossings of the sine waves or crossings of the input signals with the sine waves.

Figure 10 shows the top-level block diagram of the modulator/amplifier. Sinusoidal input test signals are generated in quadrature and drive the real and imaginary inputs of the Event Generator block, detailed in Figure 11. The event generator derives a series of 14 short pulses, or event triggers, that are used by the AM and SSB modulators. The AM and SSB modulators are detailed in Figure 12 and Figure 13, respectively.

Suppressed Carrier

It is not necessary to have a carrier tone present in the modulator output. The AM or SSB signal may have a carrier present or may be operated in a suppressed carrier fashion. When using the staggered drive we can set the DC bias level to zero so that carrier is suppressed. In the suppressed carrier case any given pulse in the tri-level waveform can be positive going or negative going, depending on the timing order of the staggered drive edges. In the suppressed carrier case, the SSB ModAmp functions as a frequency translating 'bandpass' amplifier. That is, the ModAmp frequency translates the input spectrum to some other band of frequencies as determined by the carrier frequency.

Bi-Level Alternative

For true binary output, we want the outputs to take on only two levels. The tri-level AM and SSB waveforms can be converted to bi-level waveforms by adding a square wave at the carrier frequency, as shown at the bottom of Figure 8 and Figure 9. Adding the square wave will change the carrier level of the tri-level output, however the DC bias may be changed to allow bi-level operation with or without suppressed carrier. The bi-Level output signal allows an amplifier to be built using a half bridge (instead of a full H-bridge) requiring only two output transistors. The bi-level modulators may also be operated with suppressed carrier by properly setting the DC bias.

More Simulation Results

Figure 14 to Figure 19 show several simulation results from the generalized ModAmp system. Time domain in spectrum for the following modulators are shown in the figures: Tri-level AM suppressed carrier, Tri-level SSB

suppressed carrier, Bi-level AM suppressed carrier, Bi-level SSB suppressed carrier, AM modulators with carrier, and SSB modulators with carrier.

Single Edge Modulation Alternative

The tri-level AM and SSB modulator output waveforms defined above have both leading and trailing edge modulation. An alternate approach is single edge modulation. For example, we can synthesize a tri-level AM signal with a fixed leading-edge (at the zero crossing of the sine wave) and modulate the timing of only the trailing edge. This single edge modulated AM waveform may be converted to a bi-level waveform by adding a square wave at the carrier frequency. A benefit of this waveform is that it only has 2 transitions/period (instead of 6 for standard bi-level AM). A drawback of this approach is that it has spectral distortion in the modulated output signal.

Modulator Characteristics

Table 1 summarizes the characteristics of the various modulators. We assume an H-bridge requires 4 MOSFETs and a half-bridge requires 2 MOSFETs. The "transitions per carrier period" indicate the number of signal transitions of the modulator output per carrier period. The fewer transitions generally yield higher efficiency amplifiers.

Item 5 uses two bi-level AM modulators and takes the difference to synthesize the SSB output. Item 6 starts with "bi-level AM" and adds a square wave at the 3rd harmonic of the carrier. This is used to reduce the amplitude of the 3rd harmonic in the modulator output. Schemes that add higher order harmonics is also feasible. Item 7 starts with "tri-level AM" and adds a square wave at the 3rd harmonic of the carrier. Item 8 combines two "bi-level AM reduced 3rd harmonic" (item 6) to synthesize the SSB output.

Table 1: Characteristics of the various modulators

	Modulation Technique	# of Levels seen by load	# MOSFET switches required	transitions per carrier period	(transitions/period) per switch pair	
1	Tri-Level AM	3	4	4	2	
2	Tri-Level SSB	3	4	8	4	
3	Bi-Level AM	2	2	6	6	
4	Bi-Level SSB	2	2	10	10	
5	SSB (Bi-Level AM x 2)	3	4	12	6	Two independent Bi-Level AM modulators
6	Tri-Level AM (reduced 3rd harmonic)	3	4	8	4	This is Bi-Level AM with the 3rd harmonic square wave added.
7	Bi-Level AM (reduced 3rd harmonic)	2	2	10	10	This is Tri-Level AM with the 3rd harmonic square wave added.
8	SSB (Bi Level AM reduced 3rd x 2)	3	4	20	10	

Power Supply Rejection Extension

The output voltage level of the ModAmp will be proportional to the power supply voltage unless we explicitly implement the feedforward technique suggested by equation (10). In the simulations, it was assumed that the power supply voltage was a constant voltage of 1. However, if we modify the ModAmp to monitor the power supply voltage, and make adjustments to the pulse widths, they will automatically compensate for power supply voltage variations and noise (such as the 120 Hz and other AC line harmonics).

Power supply rejection may be achieved by using a feedforward technique where the pulse-width is changed in response to a change in the power supply voltage. From (10) it can be seen that x is scaled by K/V before taking the arcsine. We have assumed that $K = 1$ and $V = 1$ in the system simulations up until now.

Figure 20(a) shows the system with power supply rejection that explicitly implements equation (10) when using the triangle wave based tri-level pulse generator.

An alternative implementation is achieved for the sine wave tri-level pulse generator case. Figure 20(b) shows a system where the amplitude of the sine wave reference is varied proportionally to the power supply voltage. To see why this works, set the sine wave amplitude parameter to

$$A = \frac{V}{K} \tag{13}$$

substitute it into (12). We get an output amplitude that is independent of the power supply voltage. Specifically, we get the desired linear relationship:

$$a_1 = K\frac{4}{\pi}x. \tag{14}$$

By using one of the feedforward power supply rejection techniques above, the usual requirement of a regulated power supply is eliminated.

ModAmp Prototype Built

A prototype ModAmp was designed and built. It performs tri-level single sideband modulation. The schematic diagram with notes is shown in Figure 21 and Figure 22. A master clock oscillator is implemented using comparator, U1. The oscillator runs at four times the carrier frequency. A simple state machine consisting of two D-flip-flops generates a pair of square waves in quadrature at the carrier frequency. Two 4th order lowpass filters are used to filter the harmonics of the square waves, leaving nearly pure sinusoidal tones. The result is the sine and cosine carrier reference signals.

The inputs to the ModAmp consists of in-phase and quadrature (Iin and Qin) audio signals. The input op-amps amplifies and hard-limits the input signals. The op-amp's output voltage is limited at the power supply rails at 0V and +5V. This limiter constrains the tri-level SSB signal's maximum pulse width. After a gain trimming pot, the audio signal is AC coupled with a 1uF capacitor. Next a DC bias is added to set the carrier level. This signal is fed into the upper comparator, U2, and an inverted copy is fed into the lower comparator, U3.

The outputs of the comparators are fed to an edge detector circuit that generates a short 250nS pulse on both the positive going and negative going comparator transitions. These "event trigger" signals are used to set and clear the A and B halves of the H-bridge.

Shown on the second page of the schematic, a complementary pair of MOSFETs are used to buffer (and invert) the event trigger pulses. The main MOSFETs are driven by a novel circuit design that use a small pulse transformer and a pair of small MOSFETs to generate the main gate drive signal. Without going into detail, the driver circuit uses the short set and clear pulses to generate bipolar gate drive signals for the main MOSFETs. The driver design avoids cross conduction (or shoot through) of the MOSFETs and operates over a wide duty cycle range.

The output load is connected between the two half-bridges to extract the SSB output. In this case, a series inductor forms the lowpass filter. The high-voltage power supply for the output stage is derived by full wave rectifying the 120V AC line voltage and filtering with a capacitor. The specifications of the ModAmp prototype are shown in Table 2.

Photographs of the ModAmp are shown in Figure 23, and the ModAmp connected to its power supply is shown in Figure 24.

Table 2: ModAmp Prototype Specifications

- Output power: 200W
- Efficiency: 95-98%
- IM Distortion: 0.4% (A 1KHz input produces 39KHz and 40KHz Outputs. Distortion measured over DC-60KHz):
- Innovative circuit design minimize parts count
- Input power source: 120VAC
- Physical size:ModAmp: 1.9 x 3.8 x 0.3 inch
  Power Supply and Matching Network: 1.9 x 3.8 x 1.2 inch

Modulator Output Spectrum

Closed form analytic expressions of the modulator output spectrum can be derived for all the modulation approaches discussed (assuming sine wave input).

Digital ModAmp Variation

Up until now, we have assumed that the ModAmps are realized with analog components such as triangle wave oscillators, sine wave oscillators, and comparators, etc. It is feasible, however, to perform all the modulation operations in the digital domain assuming we have a digital (pulse code modulation (PCM)) input signal.

A digital ModAmp can be realized as follows: (1) up-sample the input PCM waveform, (2) compare the up-sampled input to a digitally synthesized sine wave, and (3) use the comparator outputs to generate the driver signals for MOSFET power switches. The problem can be reduced to finding the zero crossing times of the oversampled or interpolated PCM waveforms. (similar to the analog event generator of Figure 11). To allow accurate calculation of zero crossing times without excessive sampling rates, a root finding method (such as the Newton method) may be used.

Once zero crossing times are calculated, digital PWM logic can generate the output waveforms. If high accuracy timing resolution is required on the edges, an extremely high clock rate would be required for a digital PWM. To alleviate the requirement for excessively high clock rates, techniques such as noise shaping may be applied to dither the timing of the edges (e.g. Delta Sigma modulation).

Polyphase Variation

Multiple ModAmps may be paralleled to reduce the output ripple voltage can increase the power. Each amplifier would be operated at a slightly advanced phase from the previous amplifier. The outputs of the "staggered phase" ModAmps would be added together through the output filter inductor, for example. With this polyphase approach, it is also possible to increase the frequency of out-of-band components, thereby reducing the post filtering requirements.

FM Modulator Variation

An FM modulator version of the ModAmp may be implemented using the same elements as the AM version with the following modifications. First, set the input to the ModAmp to a constant. This gives us a constant carrier output. Second, replace the oscillator (triangle wave or sine wave depending on the ModAmp) with a voltage controlled oscillator (VCO). Finally, use the control input of the VCO as the modulator input.

Note that it is not necessary to use an AM linearizer (since DC is used as an input) or a sine wave oscillator. We simply need to create the tri-level or bi-level waveform that will result in a carrier tone at the fundamental frequency. The VCO performs the FM modulation, the comparators generate the switching signals, and the MOSFETs switches amplify the waveform. The result is an FM ModAmp.

Other Modulation Schemes

This basic principle can be extended to other modulation schemes such as quadrature phase shift keying (QPSK), quadrature amplitude modulation (QAM), etc.

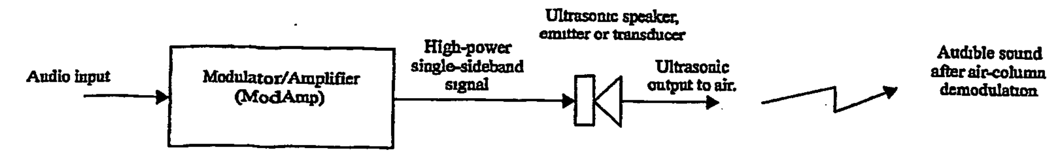


Figure 1: Use of ModAmp in parametric array application.

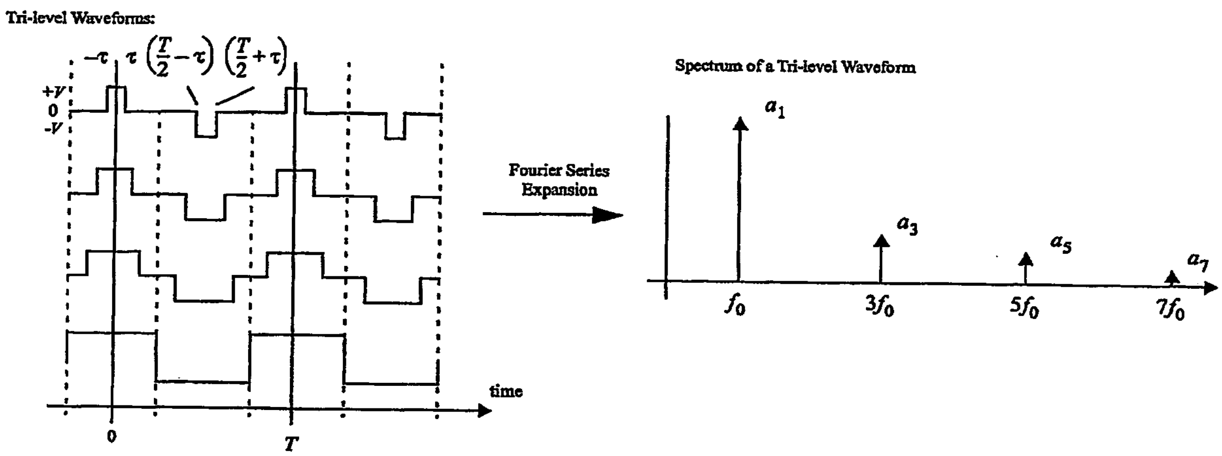


Figure 2: Bidirectional pulse waveforms with progressively wider pulse widths gives progressively stronger fundamental tone amplitudes, a_1 as given by equation (7). The spectrum of a tri-level waveform consists of the fundamental and odd harmonics.

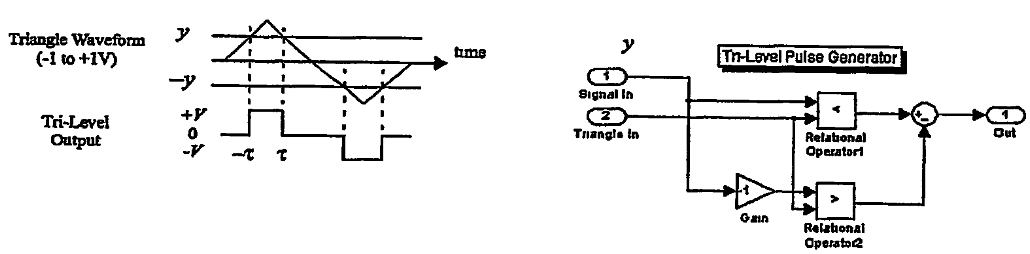


Figure 3: A triangle wave can be used to synthesize the tri-level pulse waveform. The output of the Tri-Level Pulse Generator can be directly filtered to produce a low level modulator output, or it can be amplified to deliver a high-power modulator output.

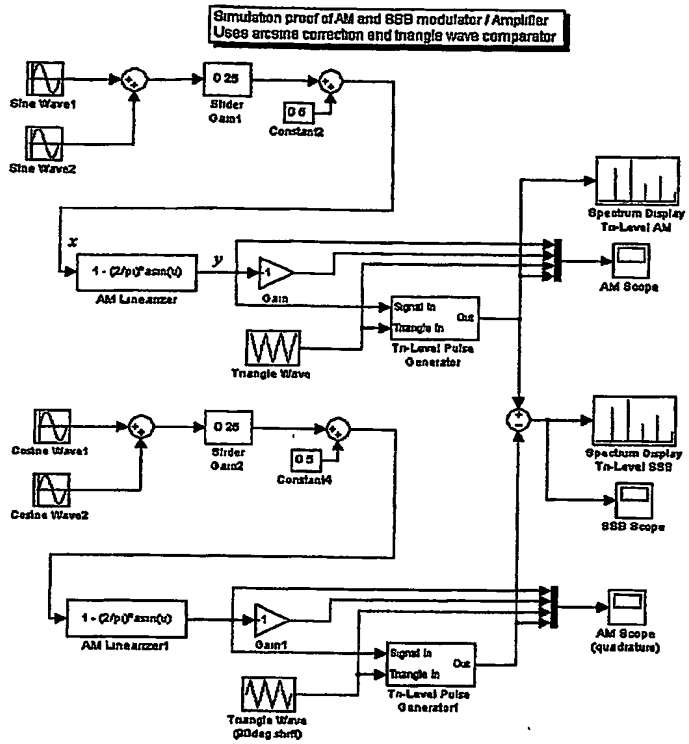


Figure 4: Block diagram of AM and SSB modulators based on arcsine linearizer and triangle wave comparator. A DC bias is added to the input signal before it is passed into the arcsine function. The Tri-Level Pulse Generator is detailed in Figure 3.

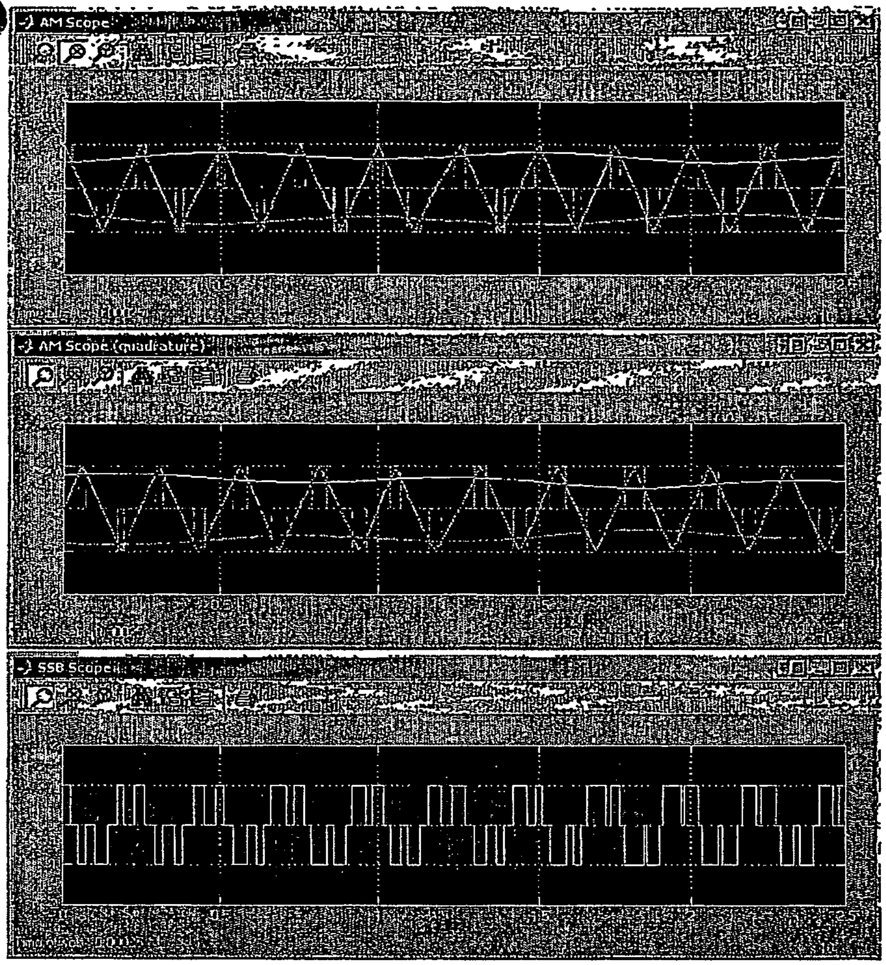


Figure 5: Time domain waveforms of simulation output. (a) The top traces show that pre-distorted audio input waveform in yellow, the inverted waveform in violet, the 40 kHz triangle waveform in blue, and the final tri-level output in red. (b) The center traces shows the AM output in quadrature. (c) The bottom trace shows the SSB output that is derived by subtracting the two AM outputs.

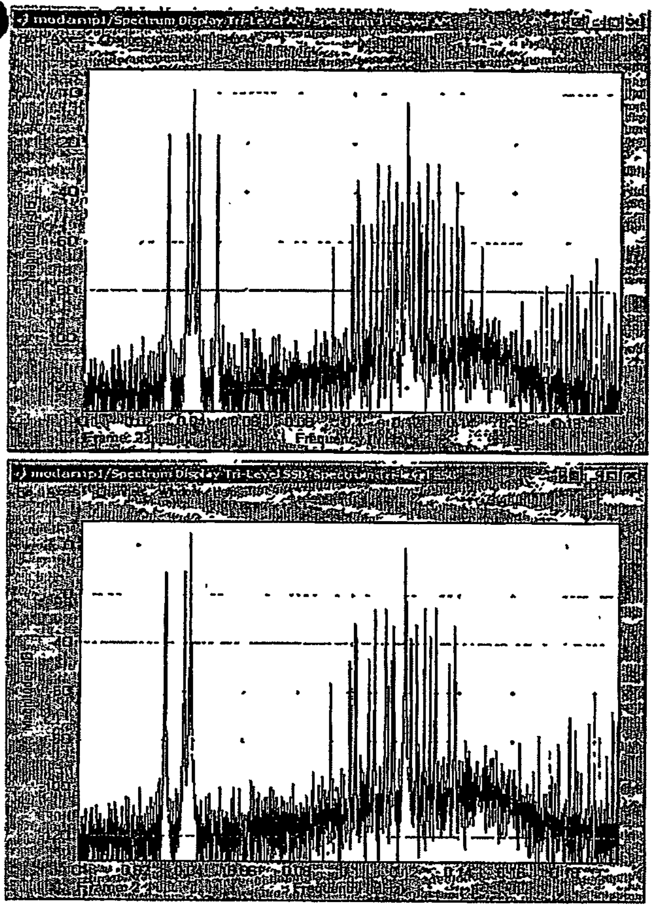


Figure 6: Frequency spectrum displays of the AM and SSB outputs. (a) The top trace shows the high spectral purity of the AM modulator around the 40 kHz carrier. (b) The bottom trace shows the spectrum of the SSB output (lower sideband).

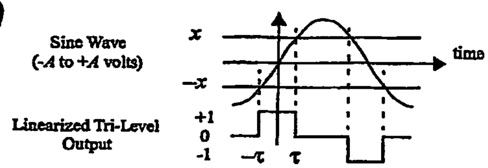


Figure 7: A sine wave can be used to directly synthesize the linearized tri-level pulse waveform and avoid the arc sine function in the signal path.

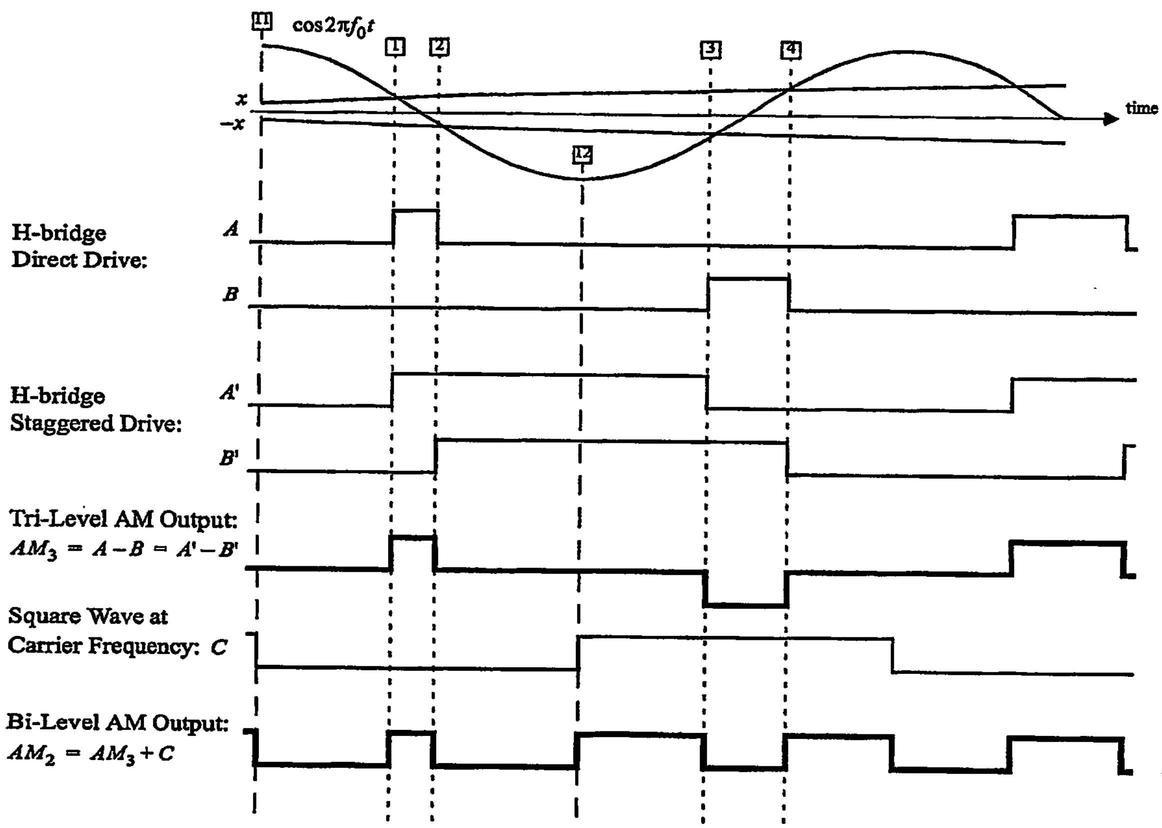


Figure 8: Sine wave synthesis of Tri-Level and Bi-Level AM. The input signal to the modulator is the x waveform. Two alternative sets of H-bridge drive signals are shown: A, B and A', B'. If the load is placed in the center of the H-bridge, the differential gives us the desired Tri-Level AM Output signal. The numbered boxes label the timing events that trigger the waveform transitions. The Bi-Level Output is derived by adding a square wave at the carrier frequency.

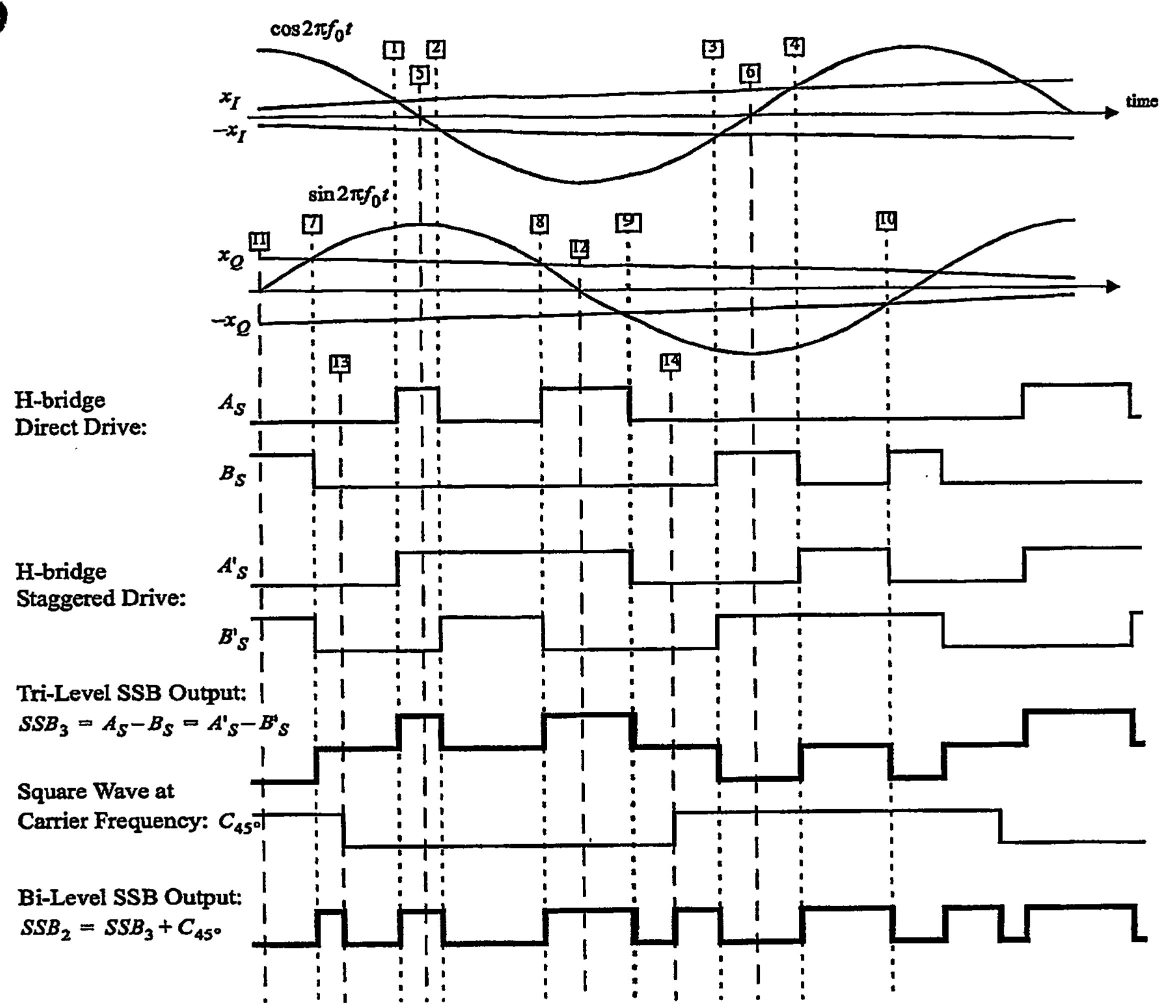


Figure 9: Sine/Cosine wave synthesis of the Tri-Level and Bi-Level SSB signals. The in-phase component of the input signal is x_I, and the 90-degree shifted quadrature component is x_Q. Two alternative sets of H-bridge drive signals are shown: A_S, B_S and A'_S, B'_S. If the load is placed in the center of the H-bridge, the differential gives us the desired Tri-Level SSB Output signal. The numbered boxes label the timing events that trigger the waveform transitions. The Bi-Level Output is derived by adding a square wave at the carrier frequency.

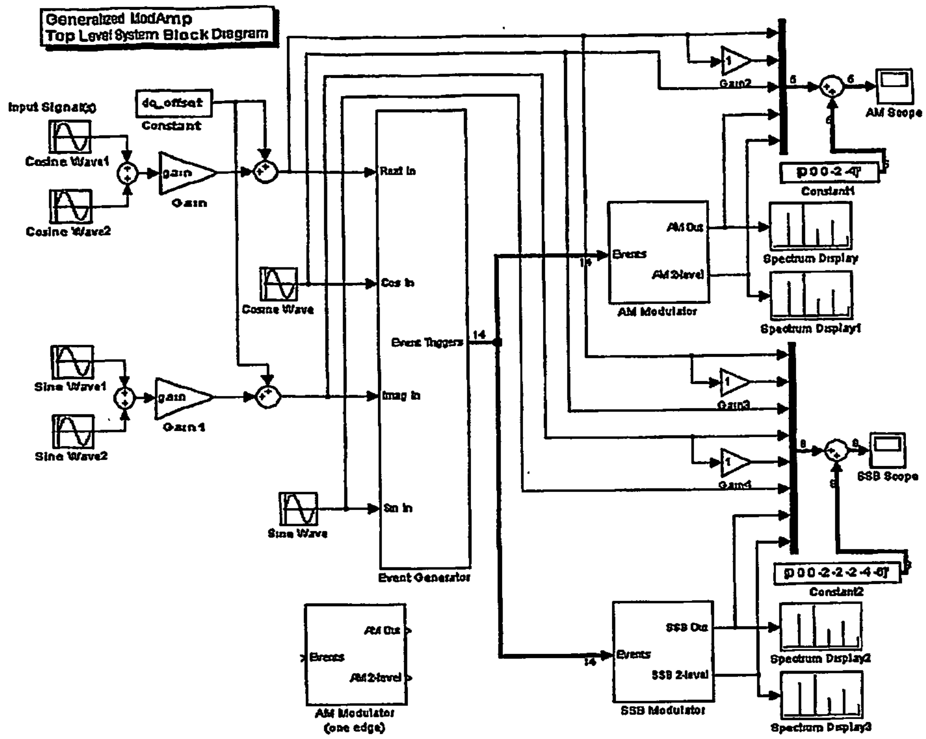


Figure 10: ModAmp top-level block diagram shows AM and SSB modulators based on Sine/Cosine wave synthesis. The Event Generator, AM Modulator, SSB Modulator blocks are detailed in the next three figures.

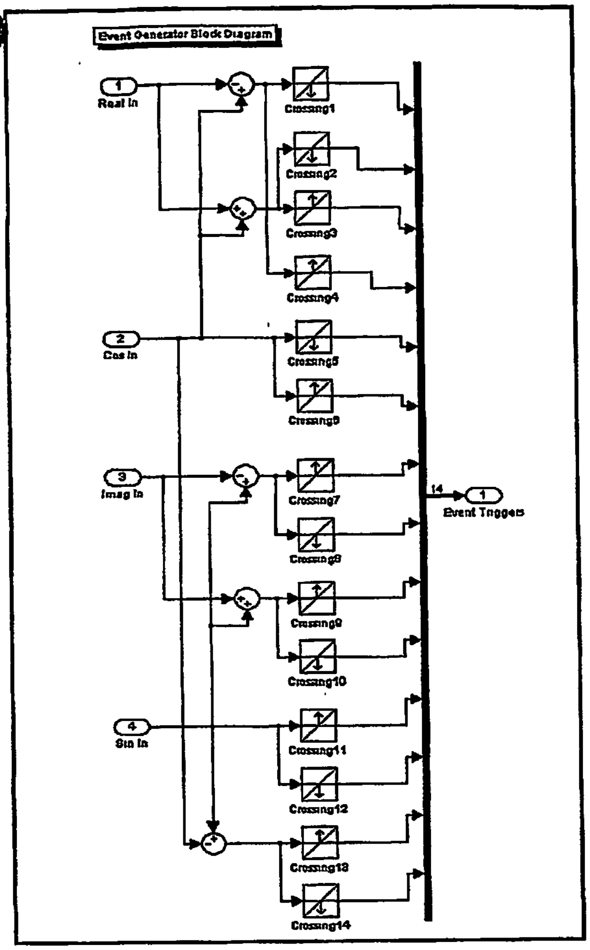


Figure 11: Event generator block diagram for ModAmp. Each zero crossing detector outputs a short pulse when the input crosses zero in the direction shown. These event trigger signals, from top to bottom, corresponding to the event numbers in Figure 8 and Figure 9.

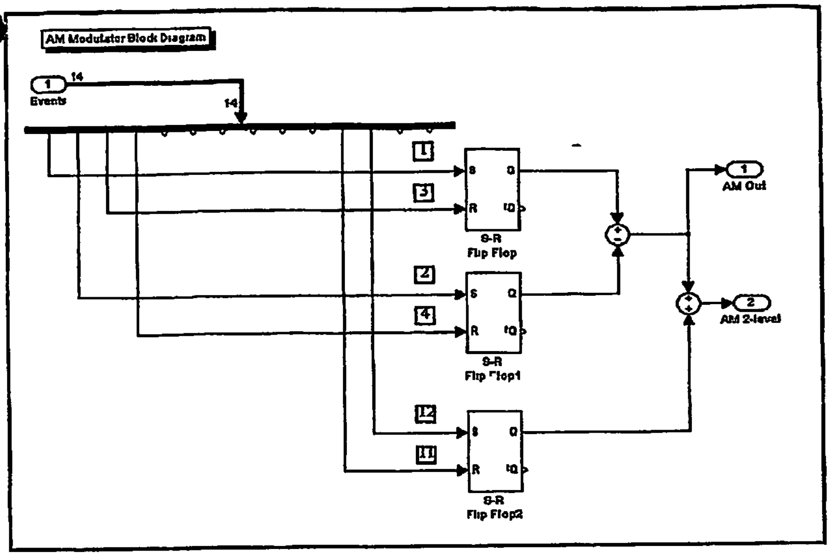


Figure 12: AM Modulator Block for Generalized ModAmp. The event triggers set and reset the filp-flops to generate the Tri-Level and Bi-Level AM outputs using staggered drive as in Figure 8.

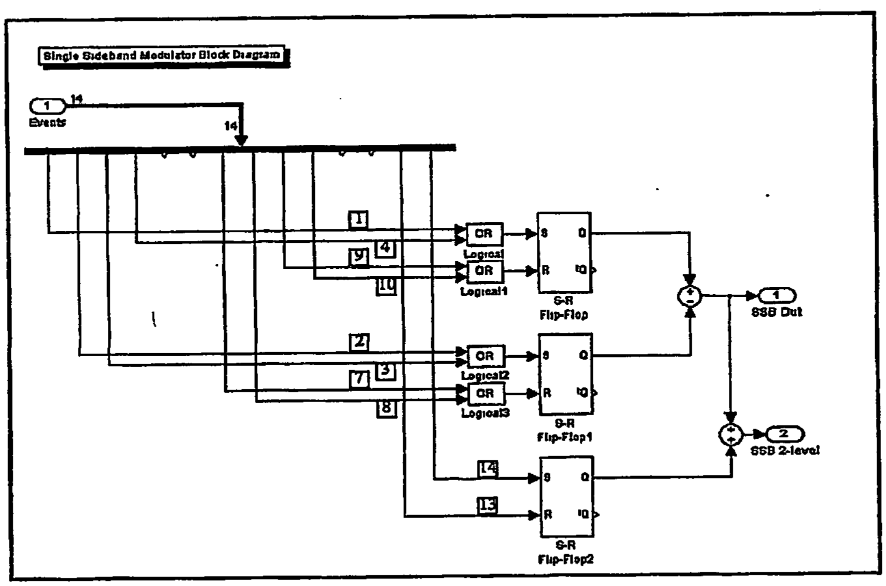


Figure 13: SSB Modulator Block for Generalized ModAmp. The event triggers set and reset the filp-flops to generate the Tri-Level and Bi-Level SSB outputs using staggered drive as in Figure 9.

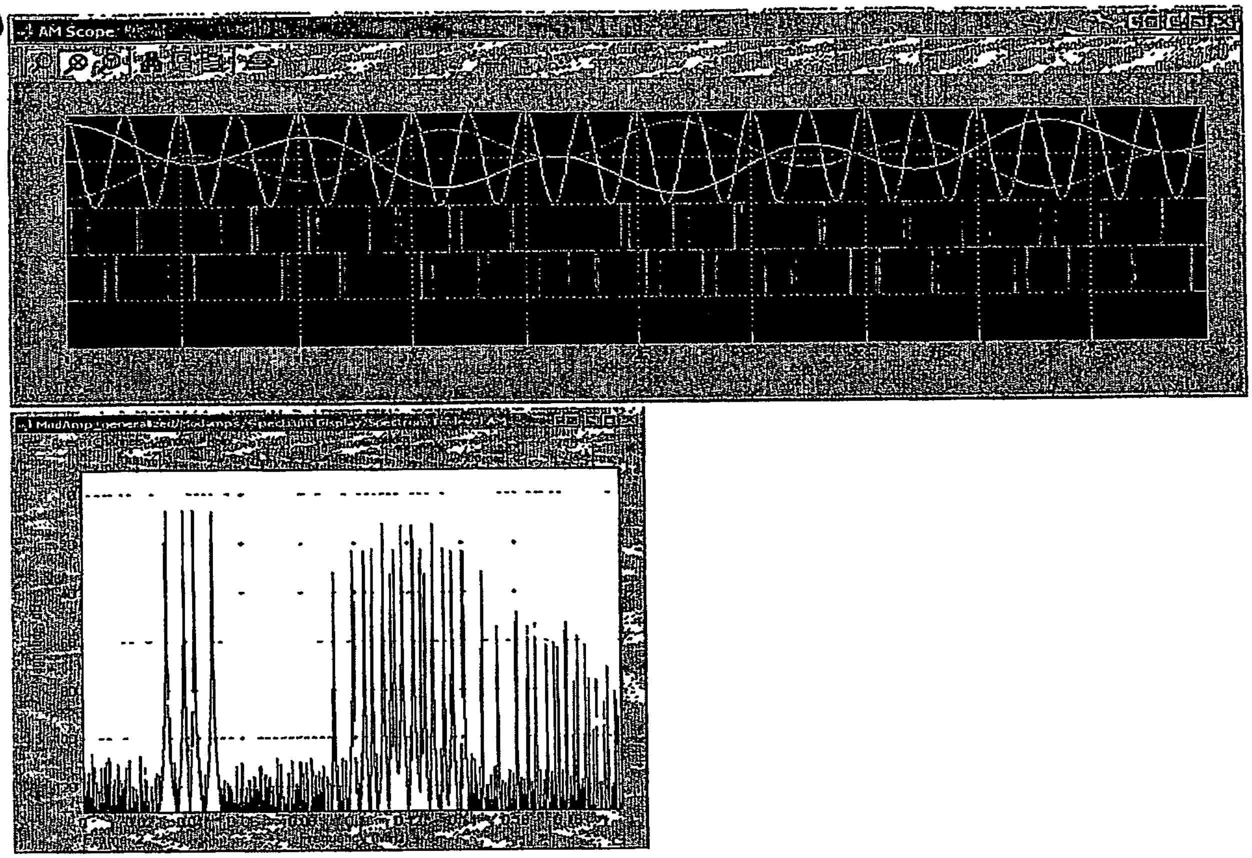


Figure 14: Tri-Level AM, Suppressed carrier. These time domain and frequency domain results were generated by the ModAmp system depicted in Figure 10. Setting the DC offset constant to zero, suppresses the carrier.

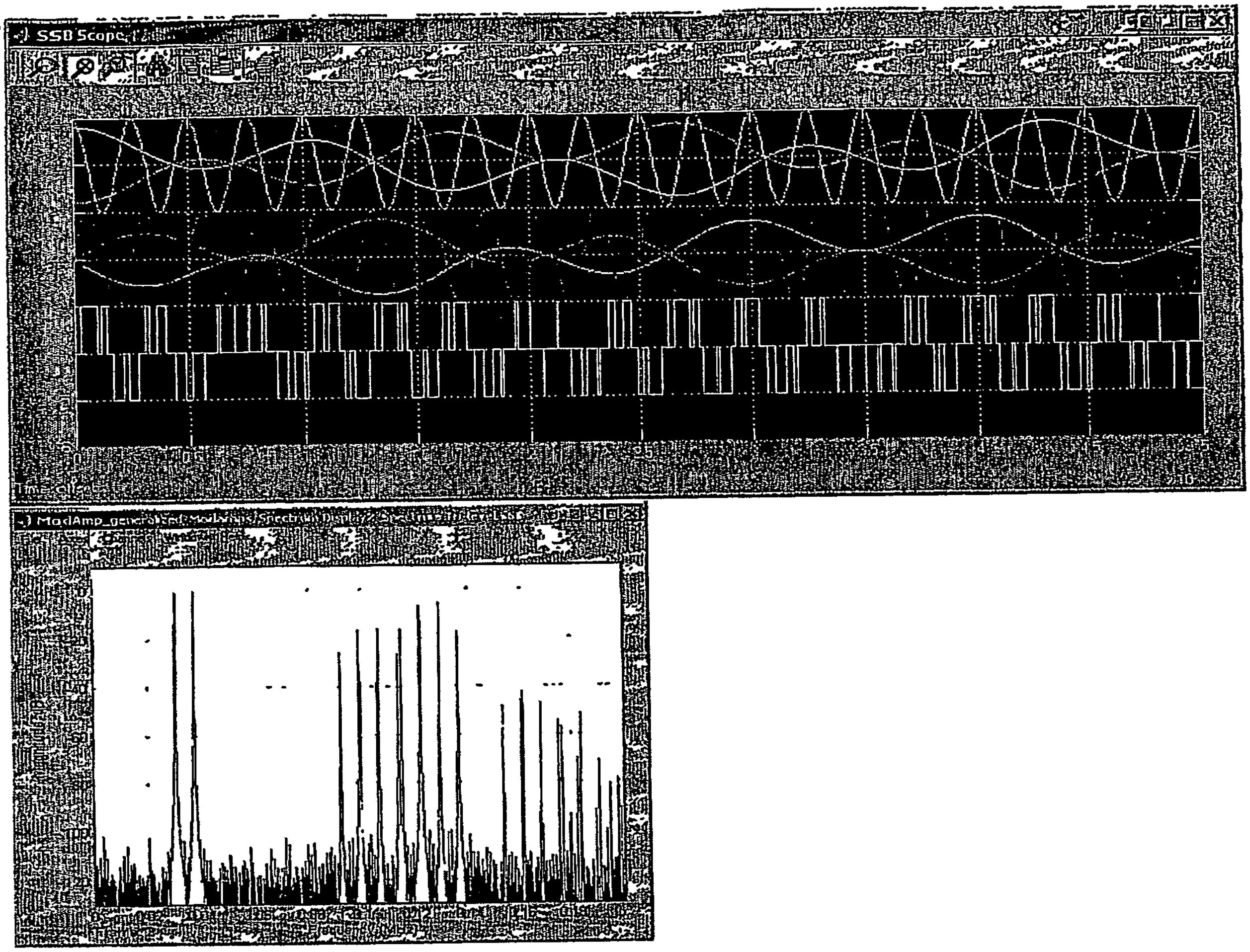


Figure 15: Tri-Level SSB, Suppressed carrier. Setting the DC offset constant to zero suppresses the carrier in this lower sideband modulator example. The time domain plot above shows the sine and cosine reference waveforms and the in-phase and quadrature input signals. The results were generated by the ModAmp system depicted in Figure 10.

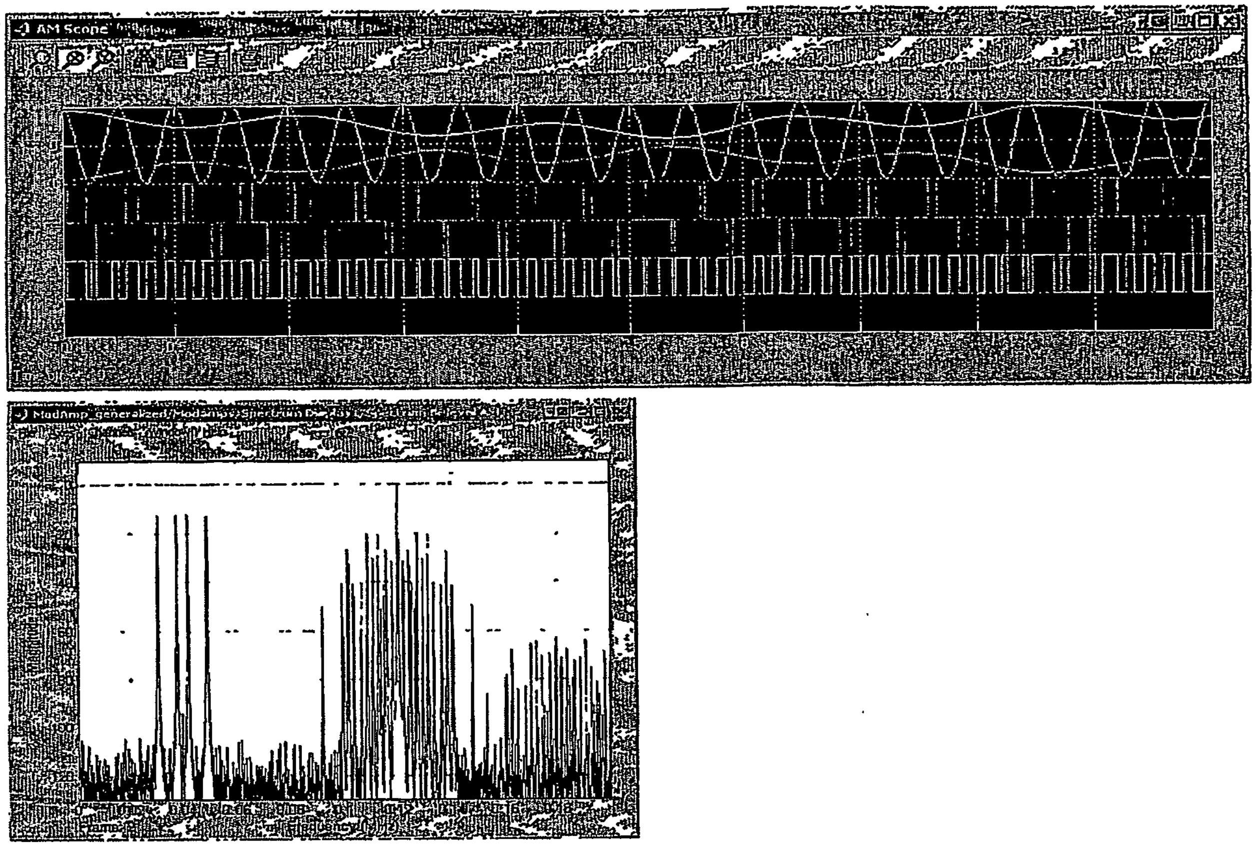


Figure 16: Bi-Level AM, Suppressed carrier. Here we set the DC offset constant to 0.5 to achieve suppressed carrier in the binary waveform (the bottom time domain waveform). The results were generated by the ModAmp system depicted in Figure 10.

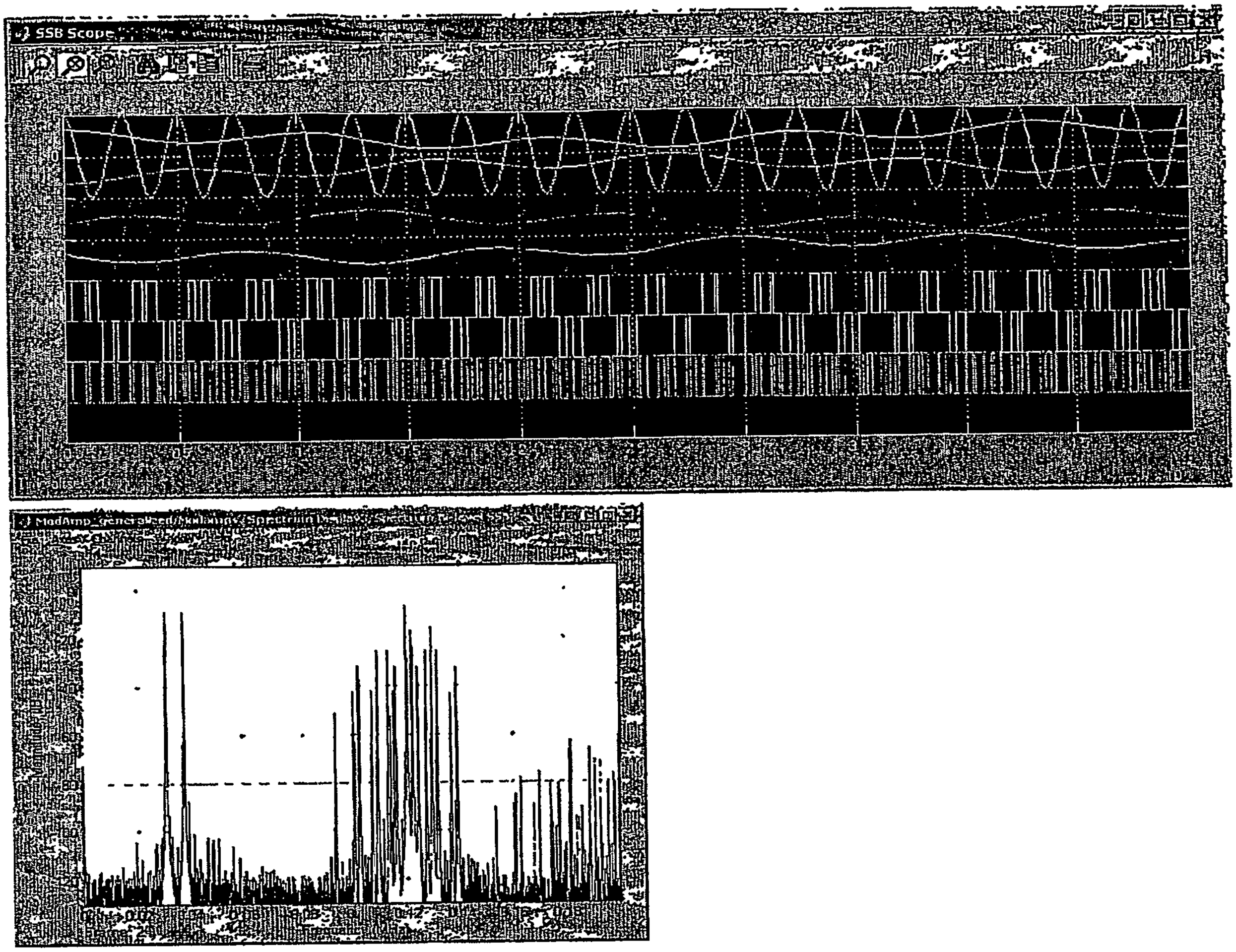


Figure 17: Bi-Level SSB, Suppressed carrier. To achieve suppressed carrier for this bi-level case, we have set the DC offset to 0.35353535. The results were generated by the ModAmp system depicted in Figure 10.

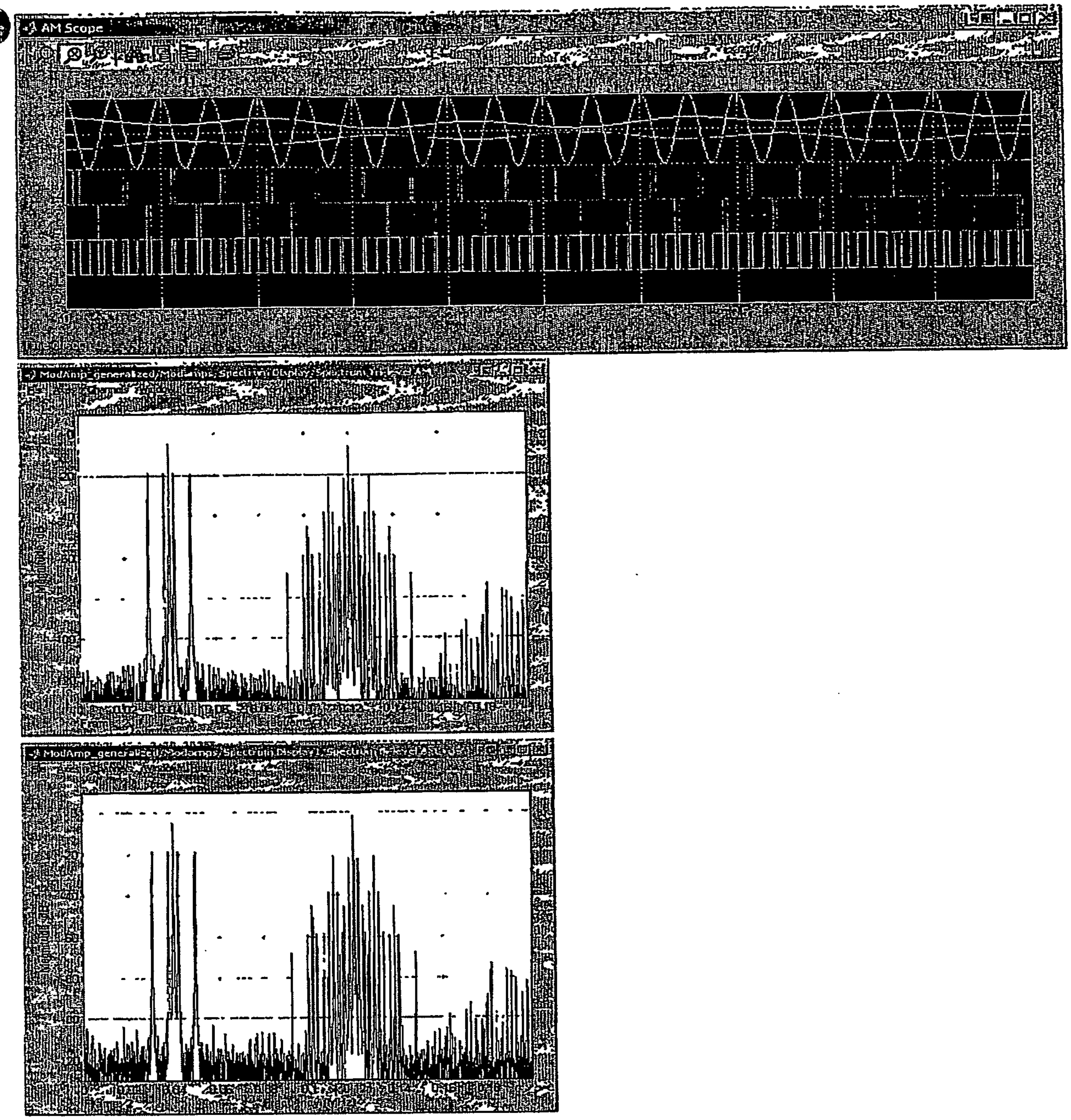


Figure 18: This simulation compares the Tri-Level (top spectrum) and Bi-Level (bottom spectrum) AM modulators (with carrier). To get the same baseband spectrum in both cases, we set the DC offset constant to 0.25. Notice that the 3rd harmonic of the carrier, at 120 kHz, is larger for the bi-level modulator. The results were generated by the ModAmp system depicted in Figure 10.

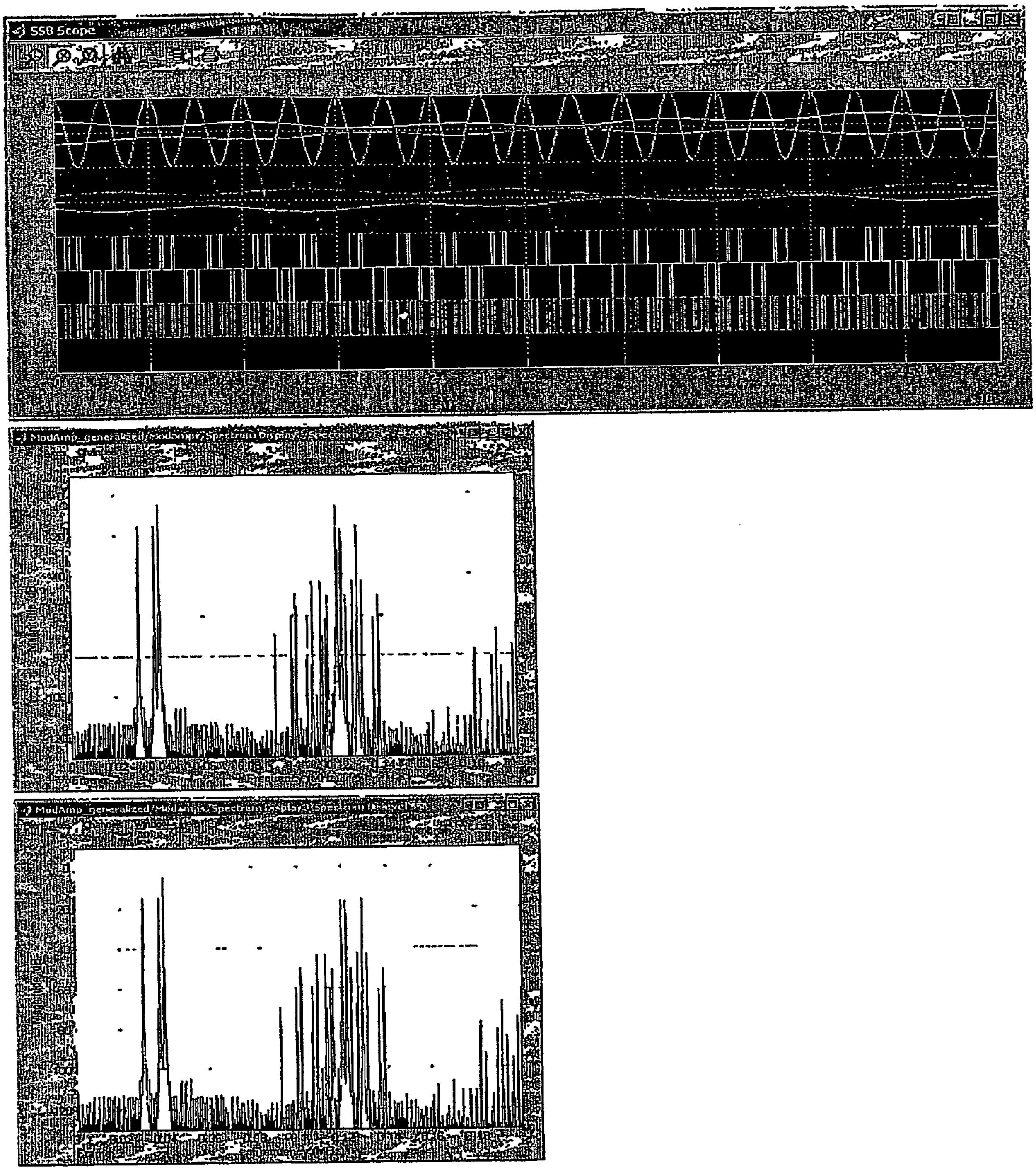


Figure 19: This simulation compares the Tri-Level (top spectrum) and Bi-Level (bottom spectrum) SSB modulators (with carrier). The DC offset constant is 0.17676767. Notice that the 3rd harmonic of the carrier, at 120 kHz, is smaller for the bi-level modulator. The results were generated by the ModAmp system depicted in Figure 10.

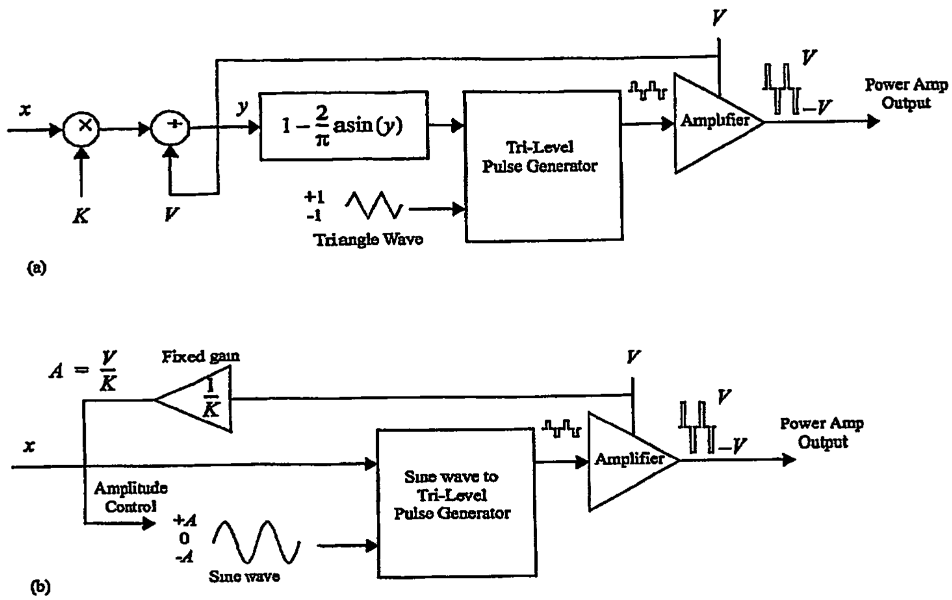


Figure 20: Power supply noise rejection may be achieved by the feedforward technique as in (a) for the triangle wave based tri-level pulse generator, or as in (b) for the sine wave based modulator. As the power supply voltage, V, changes the pulse-widths are appropriately adjusted to maintain a consistent output.

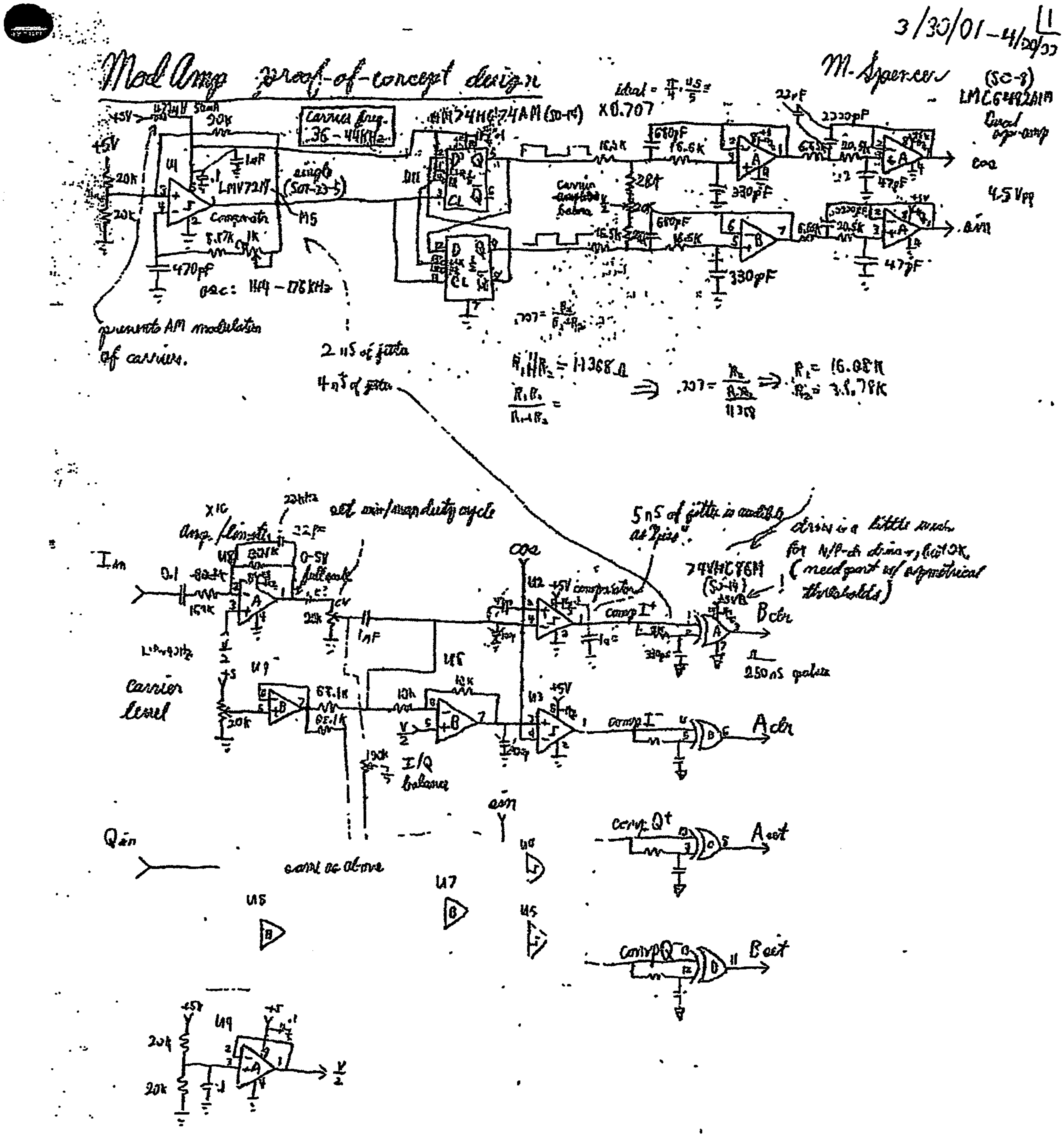


Figure 21: Schematic diagram of ModAmp prototype (page 1 of 2). See text for description.

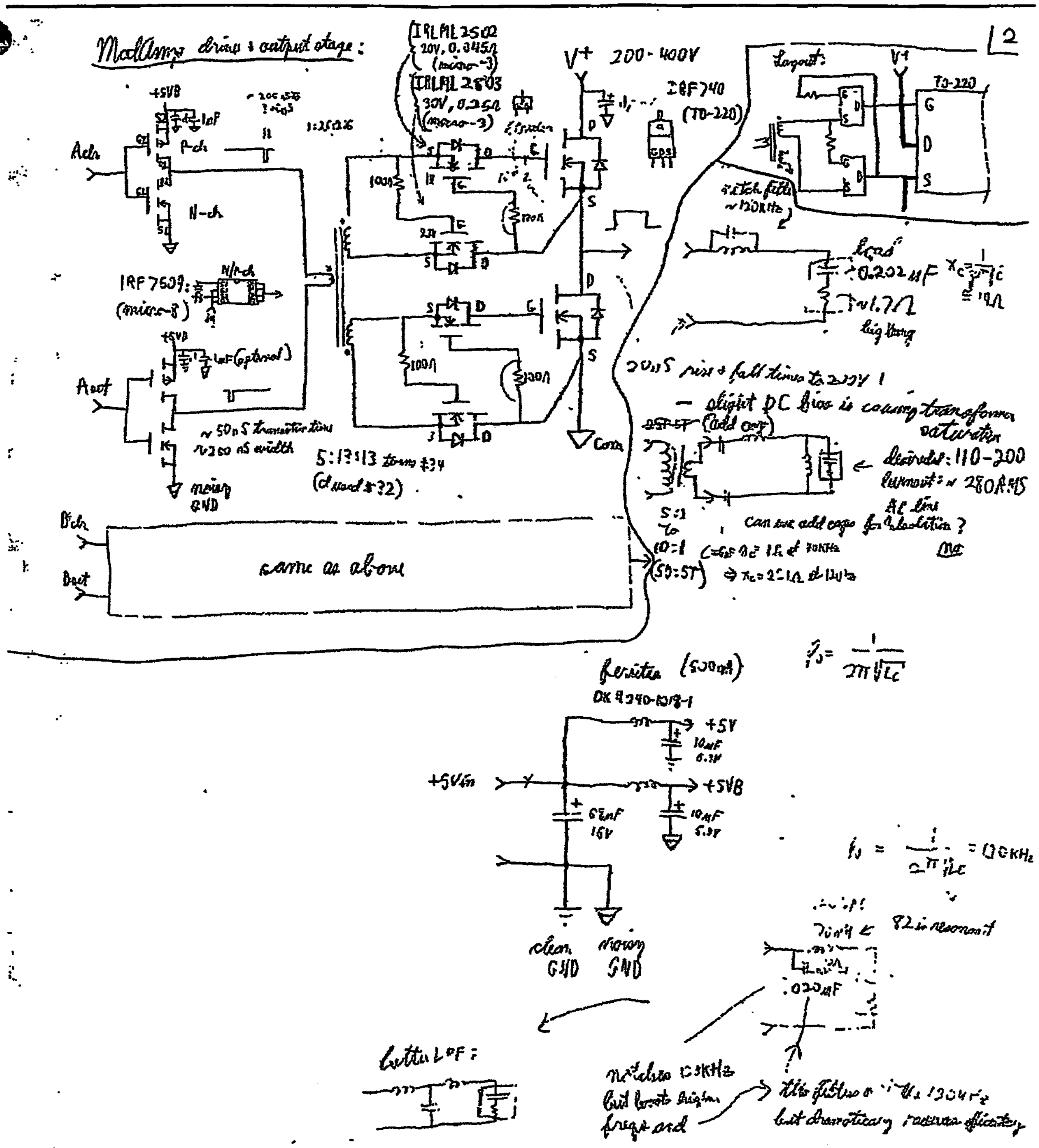


Figure 22: Schematic diagram of ModAmp prototype (page 2 of 2). See text for description.

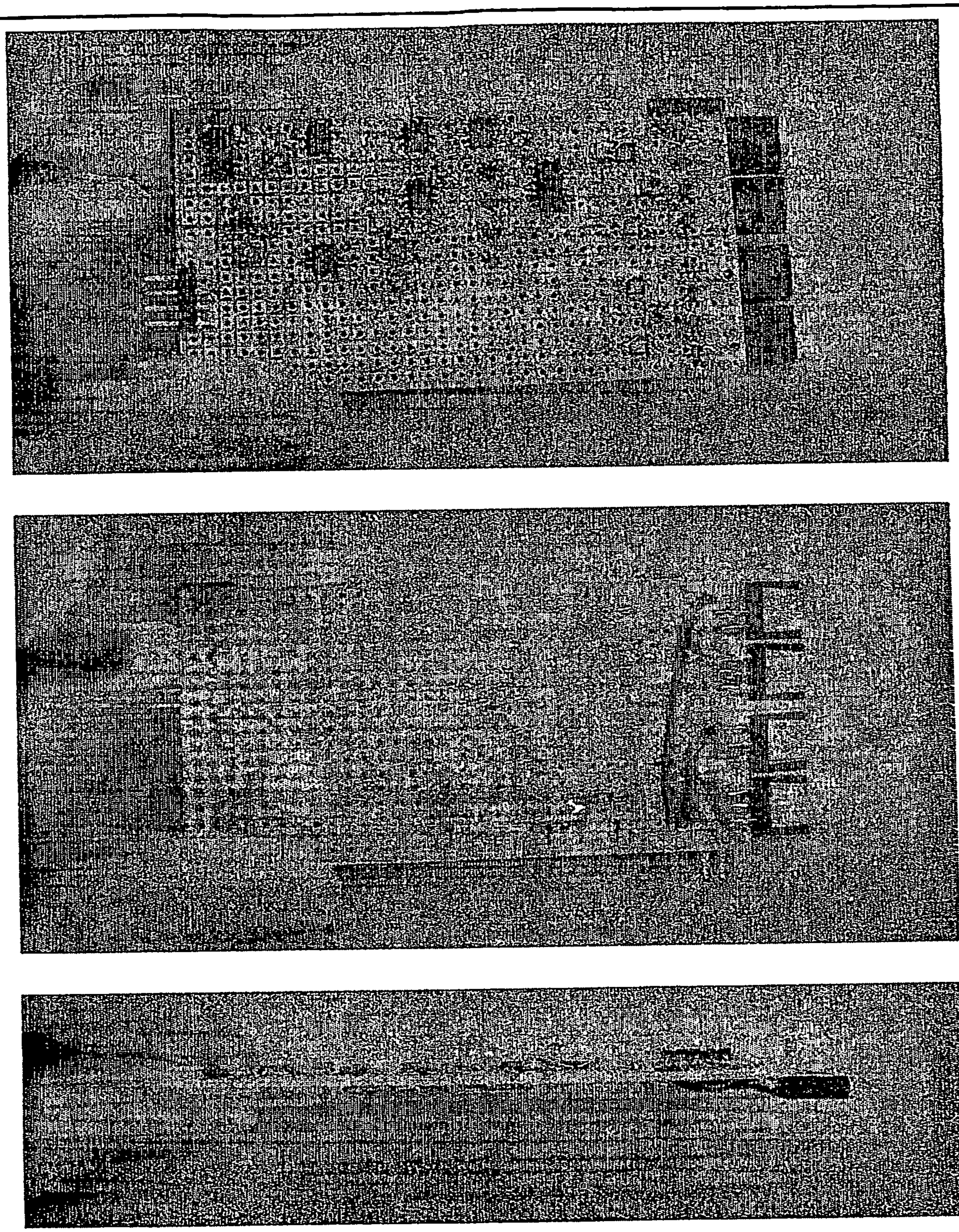


Figure 23: Photographs of functional ModAmp prototype.

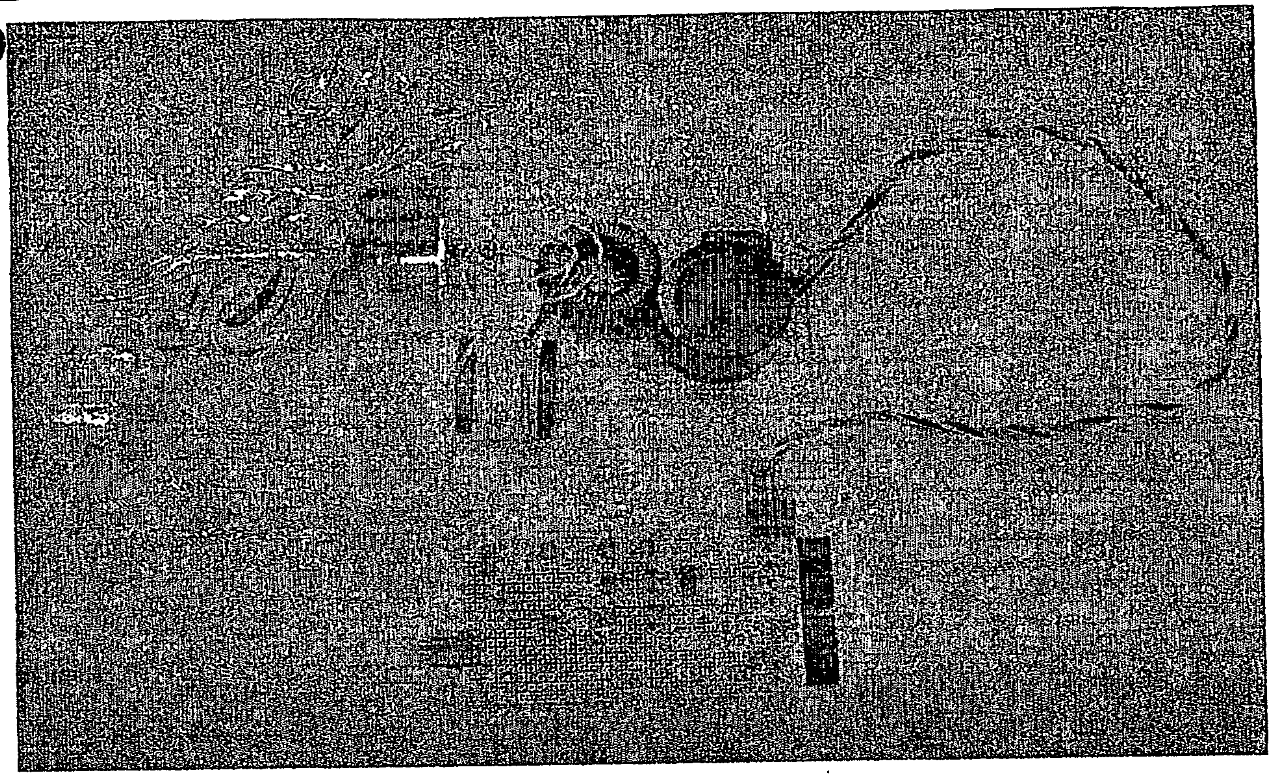


Figure 24: Photograph of ModAmp prototype (bottom board) connected to the power supply / output filter (top board).

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